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HIGH-GAIN, SELF-STEERING ANTENNA SYSTEM

ENGINEERING MODEL——DESIGN REPORT

MARCH 1966
NAS 5-10101

COMMUNICATIONS RESEARCH BRANCH
CODE 733
GODDARD SPACE FLIGHT CENTER
GREENBELT, MARYLAND

BY
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AEROSPACE GROUP
RESEARCH AND DEVELOPMENT DIVISION

HUGHES

HUGHES AIRCRAFT COMPANY
CULVER CITY, CALIFORNIA

Design Review Report

for

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PREPARED BY

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Contributors to the Design Report were personnel from the Antenna Department and the Space Electronics Department of the Hughes Research and Development Division. The following people contributed: M. Delzell, T. S. Fong, J. Howard, W. H. Kummer, F. A. Meyer, and C. R. White.

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1.0 INTRODUCTION

Previous studies and implementation of self-steering arrays have shown that high-gain antennas of this type are feasible for application to satellite communications*. The improvement in antenna gain these arrays represent, and the resulting improvement in effective radiated power, can be large when compared with antennas used on present-day communication satellites. Many other system improvements can also be achieved with these antennas.

The present program provides a logical transition between the feasibility models of the previous program and flight hardware for future applied technological satellites. A self-phasing engineering model suitable for tests in a laboratory environment will be designed, fabricated, tested, and evaluated during the course of this program. Components in this model not readily capable of being space-qualified will be specifically identified.

1.1 SELF-PHASING SYSTEM

Of the beam-steering techniques studied in the earlier program, the self-phasing arrays appear to have the most promise for future applications because they can fulfill the requirements for many earth-satellite and satellite-satellite missions. This report covers the design of one engineering model of this type and includes the appropriate trade-offs that have been made to obtain the desired objectives. Design considerations included trade-offs necessary to minimize the number of

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*Spacecraft Antenna Systems (Contract NAS 5-3545): Interim Engineering Report (Phase I - Final Report), W. H. Kummer and A. T. Villeneuve, Report No. P65-35 and Final Engineering Report, W.H. Kummer and R.A. Birgenheier, Report No. P66- , Hughes Aircraft Company, Culver City, California.

components, number of array elements, modules, radiated power per module, and primary power requirements.

A self-phasing antenna system automatically forms a high-gain beam on receive to receive a signal with a wide-modulation band. A narrow-band c-w pilot signal is used to provide an appropriately phased local oscillator for each element with which the phase of the modulation is adjusted to be equal in all the elements. The modulation, at i-f, from each channel (element) is summed, amplified, converted to r-f, amplified at r-f in a traveling-wave tube amplifier, and then distributed to each transmitting element for retransmission.

A station that desires to receive information retransmitted by a spacecraft sends up a c-w transmitting pilot signal in the up-link band. This pilot is received in the receiving antenna channels and down-converted along with the receiving modulation and receiving pilot which have been sent from an earth transmitting station. The transmitting pilot is then filtered off and sent to a transmitting channel. In the transmitting module, it is mixed with the information to obtain at each transmitting element the modulated r-f with the appropriate phase necessary to return it in the direction from which the transmitting pilot came. If r-f amplifiers are available at the frequencies involved, the inclusion at each antenna element in the transmitting antenna would considerably improve the efficiency and performance of the system.

Separate antennas are used for receiving and transmitting; these antennas are identical except that they are scaled by the ratio of the transmitting frequency to the transmitting pilot. This design serves to keep all up-link signals in the up-link band. If the transmitting pilot were allowed in the down-link band, the same antenna could be used for both transmitting and receiving.

1.2 CHARACTERISTICS OF ANTENNA SYSTEM

1.2.1

Receiving frequency	8.00 GHz ±0.175 GHz
Transmitting frequency	7.30 GHz ±0.175 GHz
Total bandwidth	Two 125 MHz information channels with a minimum guard band of 100 MHz
Polarization	Circular so that a ground station similarly equipped can receive and transmit without changing sense of polarization

1.2.2 Receiving and Transmission Functions

The system will be capable of accepting and retransmitting wide-band FM.

1.3 DESIGN OBJECTIVES

The design of the self-steered array is based on application to a gravity-gradient oriented and stabilized satellite in synchronous orbit with a cone angle of 30 degrees to allow for uncertainty in the altitude of the spacecraft. The transmission and receiving portions will steer appropriate beams along arbitrary directions within that cone. Two independent channels will be provided, and four independent beams will be steered.

A minimum gain of 30 db will be the objective in receiving and 25 dbw of effective radiated power will be the objective on transmit. The weight exclusive of power supplies and r-f local oscillator will be 110 pounds as an objective.

Attitude readouts will be provided, however, no telemetry functions will be included.

1.4 ORGANIZATION OF REPORT

The operation of the engineering model is described in detail in Section 2 of this report. Since the principles of operation of both system configurations that evolved from the design phase are basically the same, only the operation of the design selected is described. This system will serve as a communications relay for two separate communications channels, each of which operates identically. The description of the system operation is restricted to one channel.

Section 2 also presents a discussion of the weight, cost, and packaging problems that were involved in the studies of the two configurations. Attitude read-out considerations of the selected system are presented, and limitations inherent in the dynamic range of the system are given.

The signal and noise calculations and a table of the signal and noise levels throughout the engineering model system are given in Section 3. A discussion of antenna elements and array design problems and the antenna gain statistics appears in Section 4.

A specific description of electronics of the engineering model is presented in Section 5 with the aid of a numbered block diagram. Major components of the electronics system are indicated, and the specifications established for each are presented.

Detailed packaging and construction considerations involved in the design of each of the two system configurations are presented in Section 6. Section 7 contains the projected program plan.

2.0 SYSTEM DESIGN

Two approaches to the high-gain antenna system evolved during the design phase and were each studied in detail. They differ mainly in the implementation of the receiver front end. The design selected for the engineering model is called the integrated system and is shown in the block diagram of Figure 2-1. The system utilizes a wide-band receiver front end which carries the combined signals of the two communication channels until sufficient amplification has taken place to permit channel separation. This system is contrasted with the alternative approach, the dual system shown in Figure 2-2, in which channel splitting is done immediately at the receiver input with a microwave diplexer.

2.1 DESCRIPTION OF OPERATION

The principles of operation of the two system configurations which evolved from the design phase of the program are basically the same; therefore, only the integrated system design selected for the engineering model is discussed in detail. The numbers referenced in the text indicate the appropriate component on the system diagram of Figure 2-1.

The system is designed to serve as a communications relay for two independent communications channels, each with a bandwidth of 125 MHz. The frequencies allotted to each channel and the pilot signal frequencies are shown in Figure 2-1. Since the operation of both communication channels is identical, only that of Channel A is described.

2.1.1 Initial Reception

The receiving element receives an information signal at frequencies from 7.825 GHz to 7.950 GHz, a transmitting pilot at 8.001 GHz, and a receiving pilot at 7.999 GHz. These signals pass through a high-pass filter (No. 1) which is designed to pass the desired received signal and reject any signals from the transmitting antenna which may saturate the first mixer (No. 2) or which may mix with the incoming signal to produce new frequency components (cross-products) that would pass through the wide-band preamplifier (No. 5) of the system.

Following the high-pass filter (No. 1) the signal is mixed in the first mixer (No. 2) with a local oscillator (frequency 8.625 GHz) to produce the first i-f frequencies of 675 MHz to 800 MHz for the information signal, 624 MHz for the transmitting pilot, and 626 MHz for the receiving pilot. These signals then pass through the wide-band i-f preamplifier and triplexer (No. 5) which separates the two information channels (A and B) and the pilot signals. Since the first local oscillator (No. 4) is at a frequency higher than that of the incoming signals, the sense of the relative phase angles of all the incoming signals for the 64 elements are reversed.

2.1.2 Pilot Signals

Since pilot signal processing for all 64 elements is identical, the pilot signals of only one element will be considered. At the first pilot mixer (No. 6A), the pilot signals are mixed with a local oscillator frequency of 613 MHz to produce the second pilot i-f frequencies of 11 MHz for the transmitting pilot and 13 MHz for the receiving pilot. The pilot i-f's must be at a very low frequency to permit the use of very narrow-band i-f filters so that the noise in the pilot channels is reduced to a tolerable level with respect to the pilot signal.

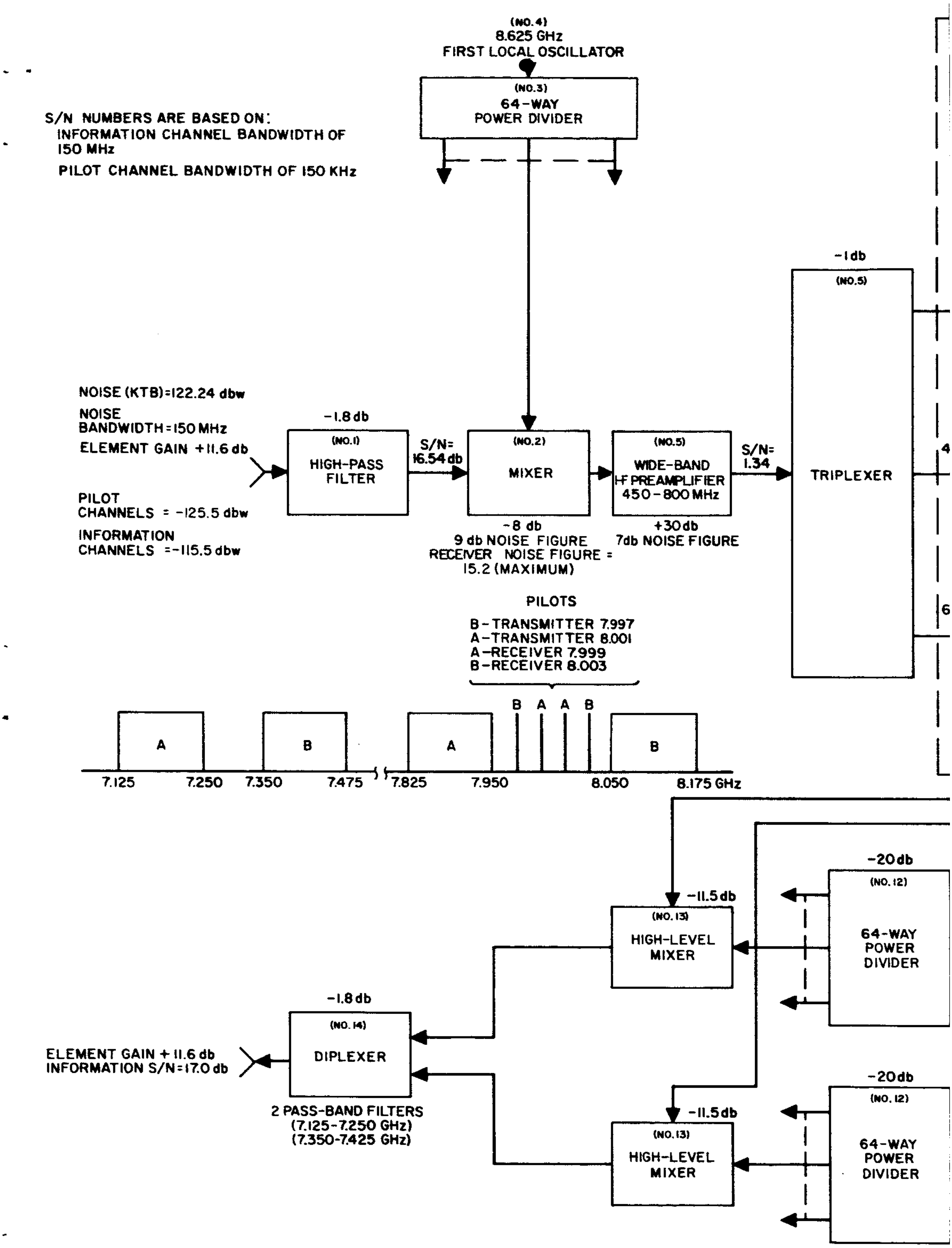
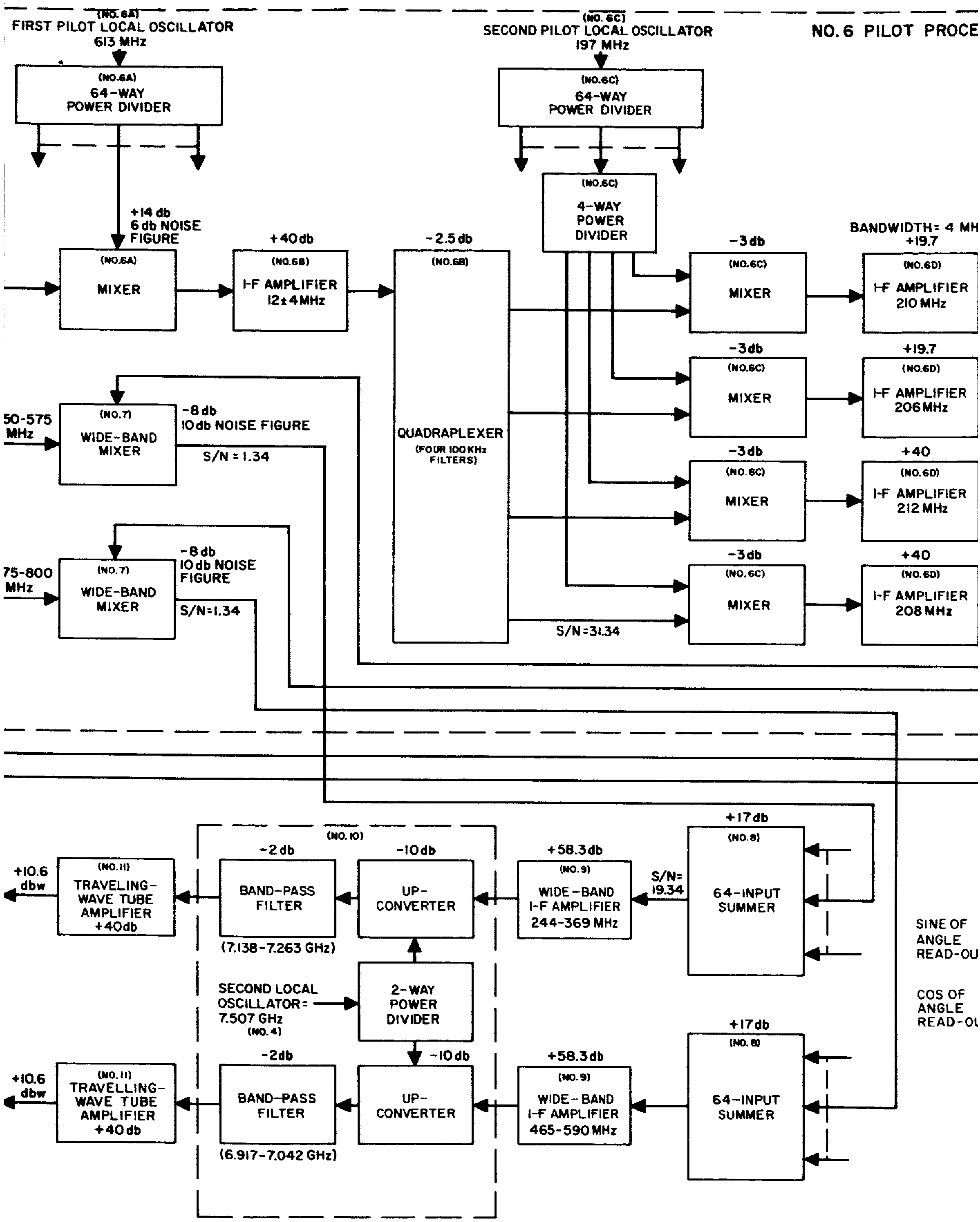
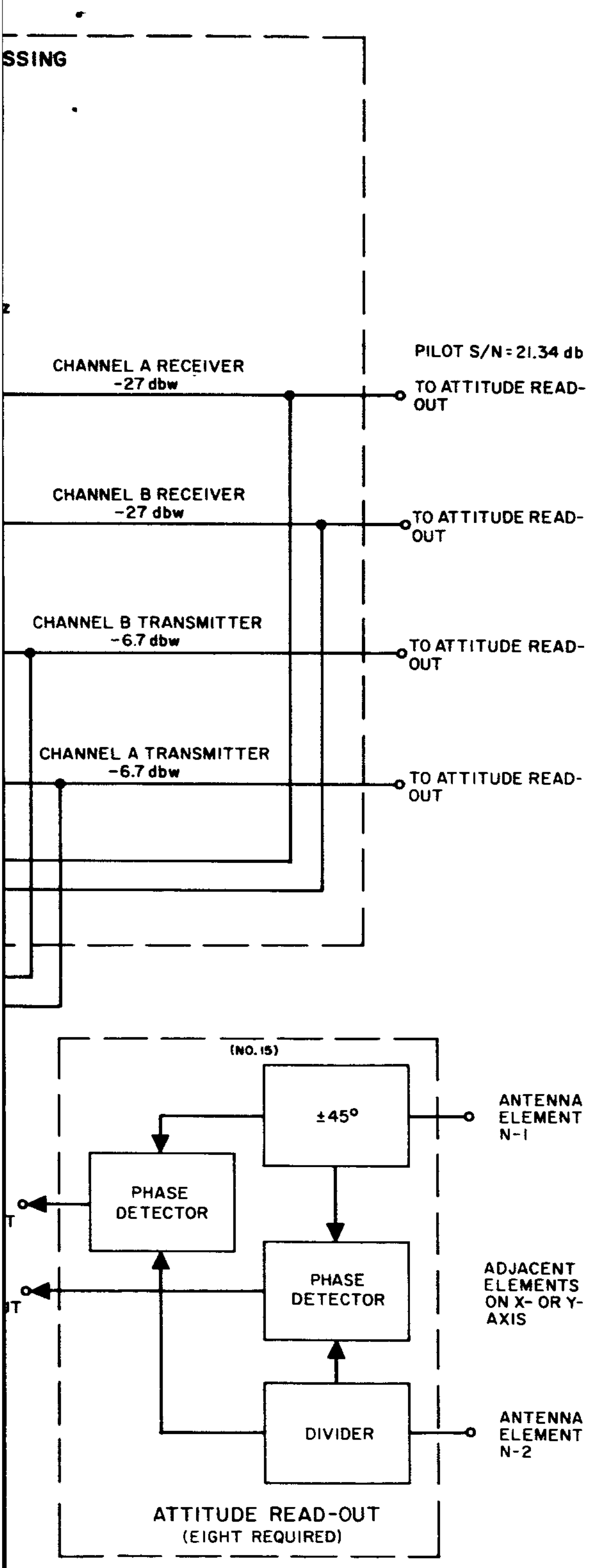


Figure 2-1. Integrated system block diagram.



NO. 6 PILOT PROCE
(NO. 6A)
FIRST PILOT LOCAL OSCILLATOR
613 MHz
(NO.6A)
64-WAY
POWER DIVIDER
+14 db
6 db NOISE
FIGURE
(NO.6A)
MIXER
+40db
(NO.6B)
I-F AMPLIFIER
12±4MHz
-2.5db
(NO.6B)
QUADRAPLEXER
(FOUR 100KHz FILTERS)
(NO. 6C)
SECOND PILOT LOCAL OSCILLATOR
197 MHz
(NO.6C)
64-WAY
POWER DIVIDER
(NO.6C)
4-WAY
POWER
DIVIDER
BANDWIDTH= 4 MH
-3db
(NO.6C)
MIXER
+19.7
(NO.6D)
I-F AMPLIFIER
210 MHz
-3db
(NO.6C)
MIXER
+19.7
(NO.6D)
I-F AMPLIFIER
206 MHz
-3db
(NO.6C)
MIXER
+40
(NO.6D)
I-F AMPLIFIER
212 MHz
-3db
(NO.6C)
MIXER
+40
(NO.6D)
I-F AMPLIFIER
208 MHz
S/N=31.34
50-575
MHz
(NO.7)
WIDE-BAND
MIXER
-8db
10db NOISE FIGURE
S/N = 1.34
75-800
MHz
(NO.7)
WIDE-BAND
MIXER
-8db
10db NOISE
FIGURE
S/N=1.34
+17db
(NO.8)
64-INPUT
SUMMER
S/N=
19.34
+58.3db
(NO.9)
WIDE-BAND
I-F AMPLIFIER
244-369 MHz
(NO.10)
-10db
UP-
CONVERTER
-2db
BAND-PASS
FILTER
(7.138-7.263 GHz)
(NO.11)
TRAVELING-
WAVE TUBE
AMPLIFIER
+40db
+10.6
dbw
SECOND LOCAL
OSCILLATOR=
7.507 GHz
(NO.4)
2-WAY
POWER
DIVIDER
-10db
UP-
CONVERTER
-2db
BAND-PASS
FILTER
(6.917-7.042 GHz)
(NO.11)
TRAVELLING-
WAVE TUBE
AMPLIFIER
+40db
+10.6
dbw
+58.3db
(NO.9)
WIDE-BAND
I-F AMPLIFIER
465-590 MHz
+17db
(NO.8)
64-INPUT
SUMMER
SINE OF
ANGLE
READ-OU
COS OF
ANGLE
READ-OU


SSING
CHANNEL A RECEIVER
-27 dbw
PILOT S/N = 21.34 db
TO ATTITUDE READ-OUT
CHANNEL B RECEIVER
-27 dbw
TO ATTITUDE READ-OUT
CHANNEL B TRANSMITTER
-6.7 dbw
TO ATTITUDE READ-OUT
CHANNEL A TRANSMITTER
-6.7 dbw
TO ATTITUDE READ-OUT
(NO. 15)
±45°
ANTENNA ELEMENT N-1
PHASE DETECTOR
PHASE DETECTOR
ADJACENT ELEMENTS ON X- OR Y-AXIS
DIVIDER
ANTENNA ELEMENT N-2
ATTITUDE READ-OUT
(EIGHT REQUIRED)

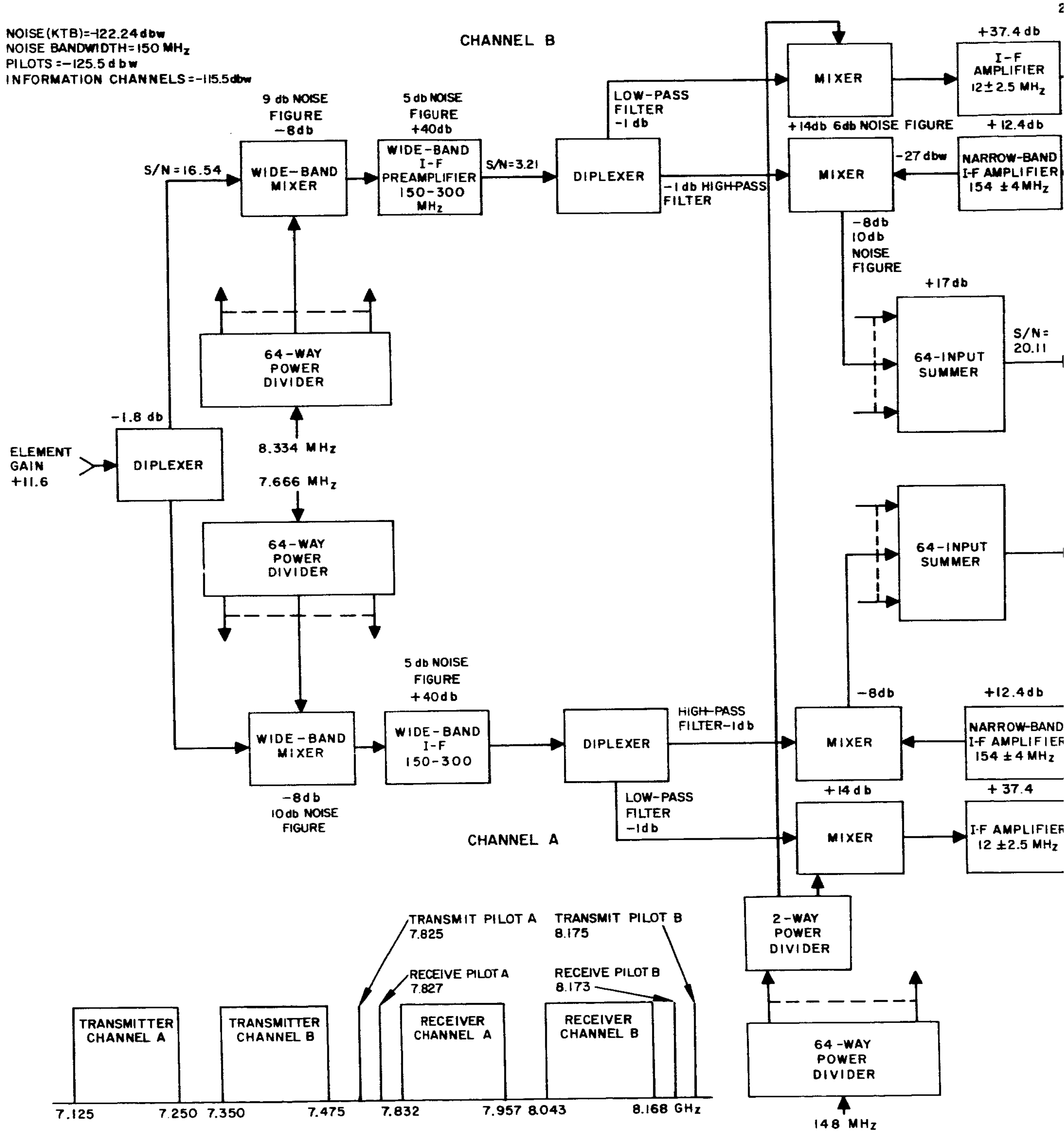
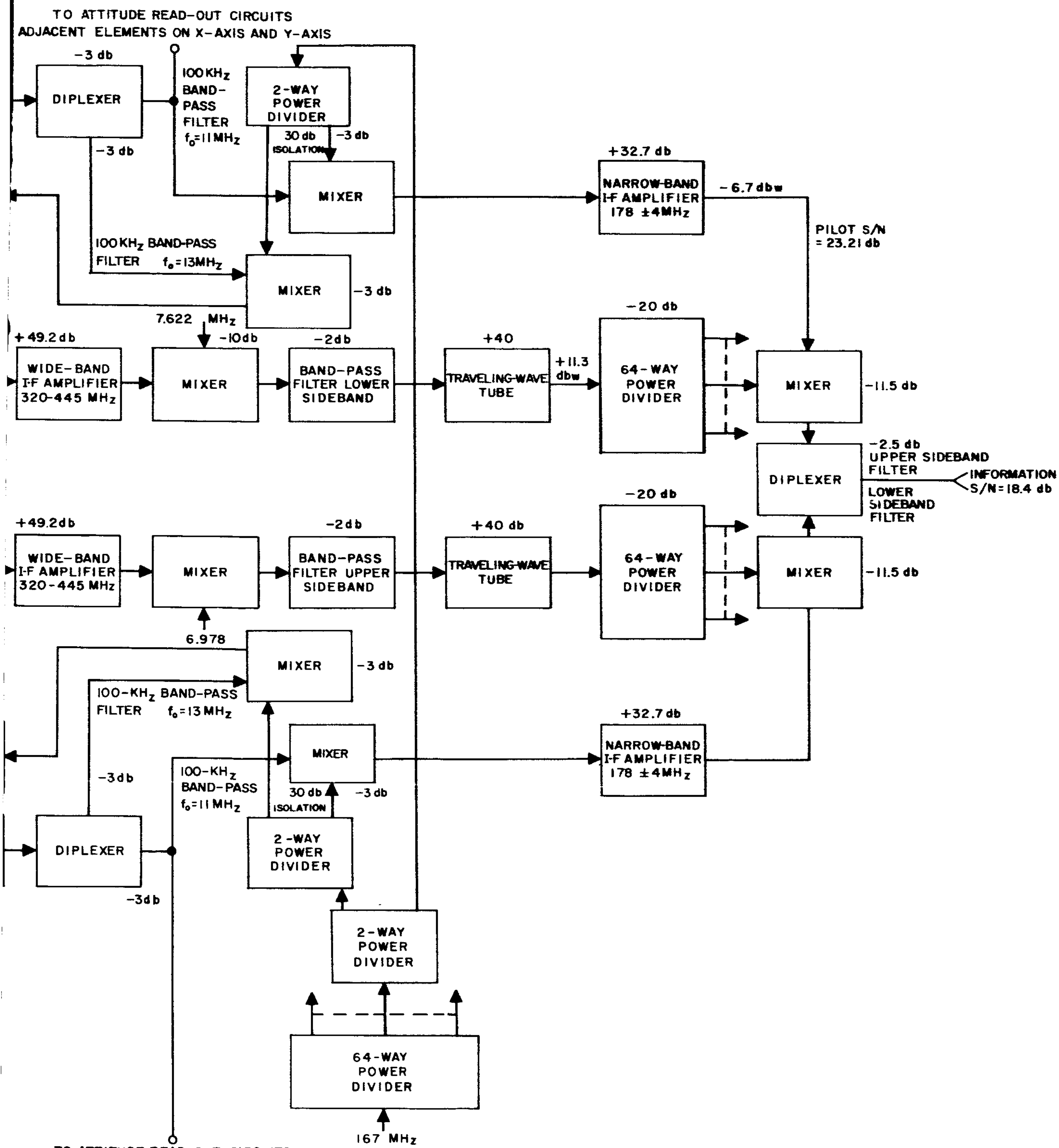


Figure 2-2. Dual system block diagram.


TO ATTITUDE READ-OUT CIRCUITS
ADJACENT ELEMENTS ON X-AXIS AND Y-AXIS
-3 db
DIPLEXER
100 KHz BAND-PASS FILTER f_o=11 MHz
2-WAY POWER DIVIDER
30 db ISOLATION
-3 db
MIXER
-3 db
100 KHz BAND-PASS FILTER f_o=13 MHz
MIXER
-3 db
7.622 MHz
-10 db
+32.7 db
NARROW-BAND I-F AMPLIFIER 178 ±4 MHz
-6.7 dbw
PILOT S/N = 23.21 db
+49.2 db
WIDE-BAND I-F AMPLIFIER 320-445 MHz
MIXER
-2 db
BAND-PASS FILTER LOWER SIDEBAND
+40
TRAVELING-WAVE TUBE
+11.3 dbw
-20 db
64-WAY POWER DIVIDER
MIXER
-11.5 db
DIPLEXER
-2.5 db
UPPER SIDEBAND FILTER
LOWER SIDEBAND FILTER
INFORMATION S/N = 18.4 db
+49.2 db
WIDE-BAND I-F AMPLIFIER 320-445 MHz
MIXER
-2 db
BAND-PASS FILTER UPPER SIDEBAND
+40 db
TRAVELING-WAVE TUBE
-20 db
64-WAY POWER DIVIDER
MIXER
-11.5 db
6.978
MIXER
-3 db
100-KHz BAND-PASS FILTER f_o=13 MHz
MIXER
+32.7 db
NARROW-BAND I-F AMPLIFIER 178 ±4 MHz
-3 db
100-KHz BAND-PASS f_o=11 MHz
30 db ISOLATION
-3 db
DIPLEXER
2-WAY POWER DIVIDER
-3 db
2-WAY POWER DIVIDER
64-WAY POWER DIVIDER
167 MHz
TO ATTITUDE READ-OUT CIRCUITS

2

After the first pilot i-f mixer, all the pilot signals are amplified by a common pilot i-f amplifier (No. 6B). The output of this pilot i-f amplifier is then directed to the quadraplexer (No. 6B), which consists of four very narrow-band i-f filters. These separate the pilot signals and establish the noise bandwidth of the pilot channels. The narrow-band filters also serve to reject the unwanted harmonics which are present at the output of the first pilot mixer.

Following the output of the quadraplexer (No. 6B), the pilot i-f frequencies are up-converted and mixed with the information signals. The frequency of the pilot signals must be at least equal to the bandwidth of the information signal to prevent overlap of the information power spectrum when the pilot and information signals are mixed. The third pilot i-f frequencies are about 210 MHz to provide about 85 MHz separation between the modulated information signal going into the high-level mixers (No. 13) and the upper sideband produced by the mixer. This separation is necessary to permit selection of only the desired sideband by the diplexer (No. 14). These third i-f pilot frequencies are produced when each pilot signal at the output of the quadraplexer (No. 6B) is directed to a separate mixer (No. 6C) in which it is mixed with a local oscillator frequency of 197 MHz.

For the transmitting pilot, the third pilot i-f frequency is 208 MHz and for the receiving pilot, 210 MHz. These signals are then directed to the final pilot i-f amplifiers (No. 6D) which serve to amplify the pilot's signal to the levels required for mixing with the information signals. These amplifiers also reject the unwanted spectral lines which have a spacing equal to the second pilot i-f frequencies.

2.1.3 Information Signals

Since the first pilot local oscillator is below the first pilot i-f frequency and the second pilot local oscillator serves to up-convert the pilot i-f frequency, the sense of the phase angles of all the pilot signals at the output of the final pilot i-f amplifiers is now the reverse of what it was at the antenna element (not considering phase shifts, which are common for all elements). Therefore, the receiving pilot has the same phase angle as the information signal at the wide-band mixer (No. 7). The information signal and receiving pilot are mixed by the wide-band mixer. The difference frequency is selected so that the phase angle of the information signal will be independent of the incoming phase angle. This mixing is done for each element of the antenna system; therefore, the information signals at the outputs of the 64 wide-band mixers (No. 7) for a channel of the system are in phase. These 64 information signals, which are at frequencies of from 465 MHz to 590 MHz, are added together in a 64-input summer (No. 8). It is at this point that the receiving array gain is realized.

The output of the 64-input summer is directed to the information i-f amplifier which has a gain of 58.3 db and serves as a band-pass filter to select the desired part of the power spectrum from the 64-input summer. Following this amplifier, the signal is mixed with a local oscillator frequency of 7.507 GHz (No. 4) and the lower side band selected by a band-pass filter (No. 10) to up-convert the information signal to a band covering 6.917 GHz to 7.042 GHz. The output of the band-pass filter is then directed to a traveling-wave tube (TWT) amplifier (No. 11) to raise the information signal to a relatively high level before distribution via a 64-way power (No. 12) divider to the 64 high-level mixers (No. 13) for this one channel of the system. If final r-f amplifiers were available, the final r-f mixing would not have to be done at a high level and the TWT amplifiers could be eliminated.

At the high-level mixers (No. 13) the transmitting pilot is mixed with the information signal and the upper sideband is selected by the diplexer (No. 14). An information signal is produced at a transmitting element with a phase angle that has the opposite sense as the phase angle of the transmitting pilot at the corresponding receiving element. The receiving and transmitting arrays are scaled in wavelength; these conditions are necessary to transmit the information from the antenna system in the direction of the transmit pilot.

2.1.4 Signal Frequencies Throughout the System

The signal frequencies for the two channels are listed in Table 2-1.

Table 2-1. Signal frequencies throughout the system.

	Channel A Frequencies (MHz)	Channel B Frequencies (MHz)
Up-link information signal	7825-7950	8050-8175
Receiving pilot	7999	8003
Transmitting pilot	8001	7997
First local oscillator	8625	8625
First information i-f	675-800	450-575
First receiving pilot i-f	626	622
First transmitting pilot i-f	624	628
First pilot local oscillator	613	613
Second receiving pilot i-f	13	9
Second transmitting pilot i-f	11	15
Second pilot local oscillator	197	197
Third receiving pilot i-f	210	206
Third transmitting pilot i-f	208	212
Second information i-f	465-590	244-369
Up-converter local oscillator	7507	7507
Up-converted information signal	6917-7042	7138-7263
Down-link information signal	7125-7250	7350-7475

2.2 ALTERNATIVE APPROACHES

The chief factors governing the selection of the integrated design of Figure 2-1 were weight and packaging, cost was also an important consideration, although it affected the decision to a lesser degree.

2.2.1 Weight and Packaging Considerations

Early in the design phase it appeared that the two systems were comparable in performance. Detailed layouts of both designs were made to provide an accurate picture of the feasibility of packaging either system and to obtain valid weight estimates. Equivalent packaging techniques were used. Both systems were significantly higher in weight than the design goal of 110 pounds. The integrated system was estimated at 175 pounds compared with a 250-pound estimate for the dual system. Since it is assumed that the design of the engineering model should be such that no major changes would be required in flight models that might be subsequently produced, weight is a very important factor in present considerations. Further work on the package layout and techniques during the development of the engineering model should result in some weight reduction in either case. However, it was felt that that design goal could only be approached with the integrated system.

The dual system results in an extremely difficult packaging problem because of the duplication of microwave components in the front end of the unit. Coaxial cables must be run from two 64-way power dividers to 128 mixers and from the mixers to preamplifiers. Limitations on the shape factor of the microwave components and the location of connectors on them add to the severity of the problem of achieving a feasible layout. In the front end of the integrated system it is only necessary to feed one set of mixers from one 64-way power divider. The number of coaxial cables required is reduced by 192. In addition, the input filter for each element is a single pass-band filter which lends itself more readily to the straight line physical design necessary for mounting behind the antenna elements.

The problem of feeding two sets of mixers from two 64-way power dividers and tieing pairs of mixers to diplexers is encountered in the integrated design as well as in the dual system since the outputs of both systems are essentially the same. However, it is felt that the performance of components in the front end more critically affects overall system performance so that the front end package should be kept as simple as possible in case the components finally selected differ from those presently contemplated. The simpler front end package also greatly facilitates assembly of the unit just because of the reduced number of interconnections that must be made.

Fabrication costs of the dual system are higher, mainly because of the duplication of high priced microwave components in the front end. Recent budgetary estimates from microwave component vendors indicated higher prices for these components than originally anticipated. Initially it was thought that the cost of the extra complexity of the dual system was small enough to be offset by the high cost of the wide-band preamplifiers needed in the integrated approach. It was not certain at that time that the preamplifiers could be achieved without resorting to expensive transistors; the 2N3570, for example, costs $100 each in large quantities and each of the 64 preamplifiers would need at least eight. This expense is contrasted with the two narrow-band and lower frequency preamplifiers in an element of the dual system, which could be designed with 99-cent transistors. However, further study has not only increased confidence in the feasibility of the wide-band preamplifier but also indicated that a less costly transistor such as the TIXM101 at $12.65 could be used.

Thus, higher microwave component prices and a lower transistor price have amounted to a cost advantage of approximately $20,000 for the integrated system, counting just the front end components alone. Further cost advantages could be accrued from the simpler fabrication and assembly effort due to the lowered cabling requirements of the integrated system.

2.2.2 Critical Components

Front end performance of the integrated system is slightly lower than that of the alternative dual-channel design. However, the difference is not serious enough to override the weight and packaging factors. The lower performance results from a lower i-f noise figure due to the higher frequency that must be used. The upper edge of the pass-band is raised by a factor of 2.5. It is imperative to use more expensive lower-noise transistors just to achieve the higher frequency of operation; the noise figure degradation in the preamplifier is limited to 2 db. The noise figures of transistorized amplifiers generally degrade about 2 db per octave as frequency of operation is raised until the transistor cut-off frequency is approached.

For the integrated system, a high-pass filter is used to reject frequencies below a certain value, $f_o - \Delta f$, and to pass frequencies from f_o to 400 MHz beyond that point. For the dual system a diplexer is used and consists of two band-pass filters. The first reject frequencies below $(f_o - \Delta f)$, passes frequencies from f_o to $(f_o + 150\ \text{MHz})$, and rejects frequencies above $(f_o + 150\ \text{MHz})$. The second filter has similar characteristics but is shifted up in frequency by 200 MHz and has skirts which are not so steep.

It turns out that the rejection characteristic at f - Δf is difficult to obtain for both the diplexer and the band-pass filter, and the wide pass-band of the high-pass filter is difficult to obtain without loss. Thus about equal insertion loss is obtained for both systems.

At the present, interdigital airline filters are being considered for this application because of their small size and weight. Performance comparisons have been made with this type of filter assumed for both the high-pass and diplexer designs; an 0.8 db higher received signal-to-noise is achievable with the dual system. A 19.3 db S/N is obtained with the present integrated system. (See Section 3.)

Recent communication from a vendor has indicated the possibility of a coaxial filter meeting the limited cross-sectional area allotted to the input filter at the expense of additional length. The critical dimension of the filter is the cross-section; it is limited by the antenna element spacing which is firmly fixed by the antenna design. The coaxial filter promises a much lower insertion loss than the interdigital filter, as low as 0.5 db compared with the 1.8 db expected with interdigital filters. Waveguide filters are also being studied to determine if a design could be evolved to fit into the space allotted. It does not appear that the dual-system diplexer could be made by other techniques than the interdigital ones. It is made up of two filters which must be run back from the antenna physically in parallel. The cross-sectional area is at least twice that of a straight single filter in any case. Therefore, it is fair to compare the performance of the integrated system if a lower-loss filter is available with that of the dual system with its relatively lossy interdigital diplexer. If 0.5 db insertion loss is actually achievable without sacrifice of the rejection requirements, the performance advantage of the dual system is eliminated. However, at present, the 1.8 db insertion loss specification is being kept in the system design until higher confidence in the claims of low loss in the coaxial filter is gained.

2.2.3 Design Risks

The dual system provides a simpler basic design with fewer risks despite its higher complexity. However, this study indicated that the wide-band design is indeed feasible. Octave-band transistor amplifiers for frequencies as high as 1000 MHz have been found, and the studies have shown that several transistor types can do the job.

Another design risk is the phase tracking of the mixers over such a wide i-f bandwidth. Not much work has been uncovered in this area. However, it is not expected that the problem would be much different if the dual system with a bandwidth reduction of two were adopted instead.

The most serious disadvantage of the integrated system is in terms of reliability. The dual system offers almost complete redundancy since the two channels can be made completely independent from input diplexer to output diplexer. Although the lower frequency local oscillators were shown as common sources in the design (to save weight), they could easily be separated. In the integrated system, a failure of the local oscillator will eliminate both channels. A component failure in the repetitive circuitry is not so serious, even if it affects both channels, since the failure of an element only degrades system performance. In the common circuitry, the two systems differ only functionally in that the dual system uses separate local oscillators. However, in a flight system, redundant oscillators will be used in the integrated system to achieve independence of the two channels. If redundant local oscillators were used in the front end with the means for failure detection and switching then available, the two systems would be comparable in reliability. It will also be necessary to design the first few stages of the power divider so that a failure cannot knock out all the elements.

The integrated system was selected chiefly because of the weight problem and the lower manufacturing cost. It seems probable that the disadvantages of this design can be circumvented in either the engineering model or in subsequent flight designs.

2.3 ATTITUDE READOUT CONSIDERATIONS

For attitude information from the satellite to be unambiguous, the angle of incidence of two pilots from two different ground stations must be known. The engineering model will have this information for all four pilots. Since any pair can be used to determine the attitude of the spacecraft, the system will have a redundancy that will greatly increase reliability of the attitude determination and will allow operation with two of the attitude read-out circuits without loss of this capability for the system. In addition, the redundancy provides a means of checking the operation of the pilots of the two earth stations that are to receive information via the satellite.

The signals for the attitude read-out are derived from two pairs of adjacent elements which lie in orthogonal planes at the center of the array (Figure 2-3). One pair of elements will lie on the X-axis and the other pair on the Y-axis, as shown in Figure 2-4.

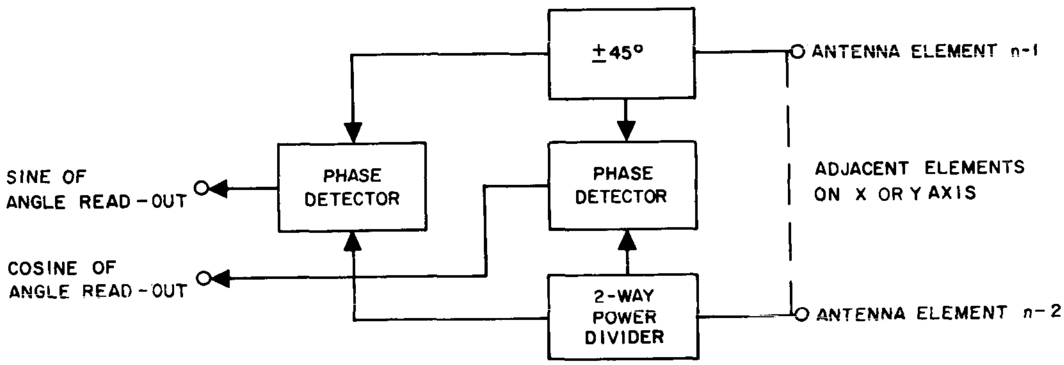


Figure 2-3. Attitude read-out block diagram.

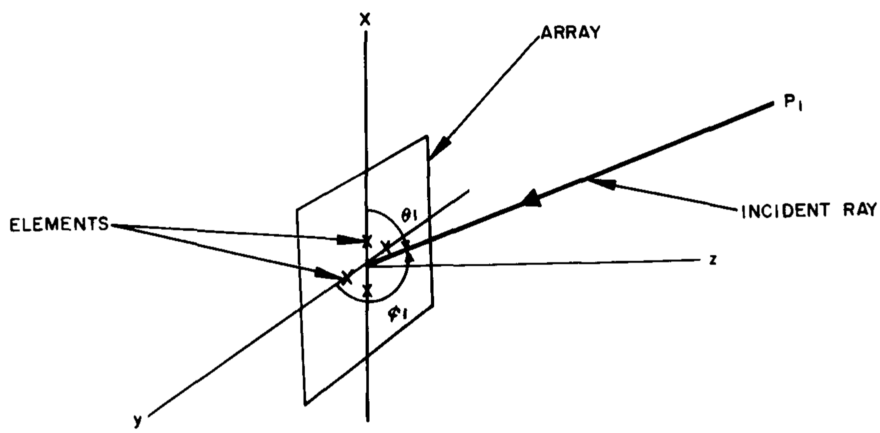


Figure 2-4. Arrangement of elements and angles determined for attitude read-out.

Signals for the angle read-out are directed from the final pilot i-f amplifiers to the attitude read-out phase detectors. Since the element spacing will be on the order of two wavelengths, the output from each phase detector will change by 360 degrees when a pilot position changes by 30 degrees;

therefore, unique specification of an angle requires quadrature phase detectors for each angle.

The output voltages from these detectors provide for a single angle as shown in Figure 2-4, and normalized to unity magnitude, are cos (Kd cos θ_i) and sin (Kd cos θ_i). If θ_i varies by 30 degrees, the maximum element separation which can be used with this type of angle read-out is d = 1.9315T. For this value, the normalized output voltages from one of the quadrature phase detectors become cos (3.863π cos θ_i) and sin (3.863π cos θ_i). These voltages are plotted as a function of θ_i from $\theta_i = 75^\circ$ to $\theta_i = 105^\circ$ in Figure 2-5.

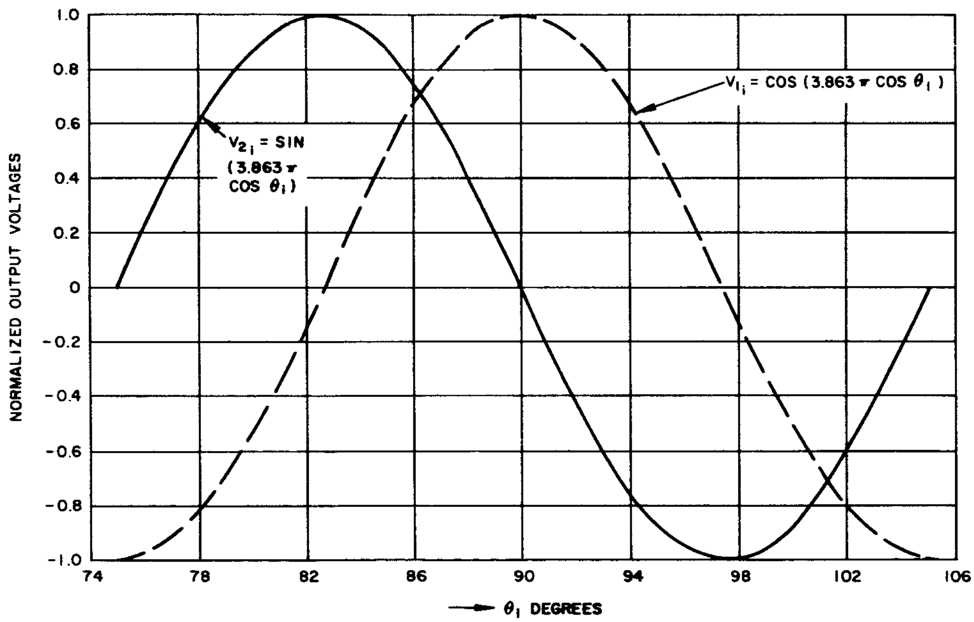


Figure 2-5. Attitude read-out voltage for angle θ_i.

The design objective for the accuracy of the attitude read-out is ±0.5 degree. Achieving this value means that the relative phase of the electronics cannot vary more than ±5 degrees. The effects of mutual coupling have been ignored, but the attitude read-out could be calibrated to account for them after the system had been fabricated.

2.4 DYNAMIC RANGE

Because of various factors which affect the power density of the signals received by the satellite, the communications system must have some dynamic range. However, it is also desirable that the signal levels into the high-level mixers be maintained at a constant level to realize the most efficient frequency conversion and to eliminate the necessity of making the mixers track over a dynamic range. Also, similarly, the receiving pilot signal levels should be maintained at a constant level to realize the optimum noise figure and conversion efficiency for the wide-band mixers. To maintain these signals at a constant level, the final pilot i-f amplifiers are limiting amplifiers and the wide-band information i-f amplifiers incorporate an automatic gain control (AGC) loop.

Several factors influence the dynamic range of the system:

1. Difference of path loss from maximum distance to minimum distance on earth = 1.3 db
2. Rain attenuation ≈ 0.2 db
3. Element factor fall-off = 3 db
4. Miscellaneous losses = 1.5 db
5. Additional margin = 2 db

The nominal values for signal levels, given for this system in Section 3.2 are for minimum element gain and maximum distance to the earth; therefore difference in path loss and element factor fall-off will increase the signal level above the values given while miscellaneous losses and rain attenuation will decrease it. Thus the signal level at the satellite can increase by 6.3 db and decrease by 3.7 db from the nominal values given. The system will be designed with this dynamic range capability.

3.0 SIGNAL AND NOISE LEVELS

The signal levels for the engineering model were determined on the basis of the required signal-to-noise ratio of the signal to be transmitted from the system and on a reasonable transmitter power and antenna size for the ground station. For a 15-foot parabolic antenna with a transmitter power (information band) of 40 dbw for a ground station, the power received at a satellite in synchronous orbit is -115.5 dbw (isotropic). In the design of the system this received power level will give a signal-to-noise ratio of 19.34 db for the information signals at the output of the wide-band i-f amplifier (9)*.

To prevent degradation of the information signals when they are mixed with the pilot signals, the signal-to-noise ratios of the pilot signals must be reasonably high; yet the power of the pilot signal should be small compared with the total power transmitted from the ground. Because of these requirements the bandwidth of the pilot channels was made reasonably narrow. The noise bandwidth of the pilot channels is 150 KHz which is the minimum practical bandwidth obtainable without resorting to very expensive crystal filters.

The effective radiated power (ERP) is the maximum which is practical to achieve without the use of final r-f amplifiers for each element which are presently beyond the state of the art. ERP for this system is 25 dbw.

---

*Numbers in parenthesis refer to the block diagram of Figures 2-1 or 5-1.

3.1 SIGNAL AND NOISE CALCULATIONS

1. Noise level at output of high-pass filter = KTB

2. Noise level at output of wide-band mixer = KTB + 1 db
   = 1.259 KTB

3. Noise level at output of wide-band preamplifier
   = (1.259 KTB) G + G (F-1) KTB
   = GKTB (1.259 + 5-1)
   = GKTB (5.259)
   = KTB + 30 db + 7.2 db = KTB + 37.2 db

   Because of the 30-db gain of the i-f preamplifier, the signal-to-noise level of the pilot channels is determined by the signal-to-noise ratio at the output of the preamplifier based on a 150 KHz bandwidth. The pilot signal level at the satellite is -125.5 dbw, which requires a pilot transmitter power of 30 dbw using the 15-foot reflector.

4. Noise level of the pilot signals at the preamplifier output based on 150 KHz noise bandwidth
   = KTB + 37.2 db
   = -204 dbw + 51.76 db + 37.2 db
   = -115.04 dbw

5. Pilot signal level at preamplifier output
   = -125.5 dbw + 11.6 db - 1.8 db - 8 db + 30 db
   = -93.7 dbw

6. Signal-to-noise ratio of pilot signals
   = 115.04 db - 93.7 db = 21.34 db

7. Noise level of information signals at the preamplifier output based on 150 MHz noise bandwidth
   = -85.04 dbw

8. Information signal level at preamplifier output
$= -83.7$ dbw

9. Signal-to-noise ratio of information signal at pre-amplifier output based on 150 MHz bandwidth
$= 1.34$ db

10. Noise level at output of 64-input summer
$= -85.04$ dbw - 1 db - 8 db - 1 db = -95.04 dbw

11. Signal level at output of 64-input summer
$= -83.7$ dbw - 1 db - 8 db + 17 db = -75.7 dbw

12. Signal-to-noise ratio at 64-input summer output
$= 95.04$ db - 75.7 db = 19.34 db

These calculations determine the signal to noise ratio for the information signal up to the high-level mixer. At the high-level mixer, the signal-to-noise ratio of the information signal is 19.34 db and the signal-to-noise ratio of the pilot which serves as the local oscillator is 21.34 db. It can be shown that the signal-to-noise ratio at the output of the high-level mixer is given by the expression

$$\left(\frac{S}{N}\right)_{out} = \frac{\left(\frac{S}{N}\right)_{in}}{1 + \frac{1}{\left(\frac{S}{N}\right)_{lo}}\left(1 + \left(\frac{S}{N}\right)_{in}\right)} \tag{3-1}$$

For the case at hand, this expression yields a signal-to-noise ratio at the output of the high level mixer of 17.0 db.

The signal-to-noise ratio of the radiated signal is higher than 17.0 db because of the incoherent addition of noise in the far field. In the far field the signal-to-noise ratio would be 19.3 db.

3.2 SUMMARY TABLE

All signal levels, component gains, and signal-to-noise ratios are listed in Table 3-1 and are included in the block diagram of the system (Figure 2-1).

Parameter	Signal Levels (dbw)	Component Gain (db)	Signal-to Noise Ratios(db)
Information Signal			
Received signal level (isotropic)	-115.5		
Receiving antenna element gain		11.6	
High-pass filter		-1.8	
First mixer		-8	
Wide-band i-f preamplifier		30	
Signal at preamplifier output	-83.7		1.34
Triplexer		-1	
Wide-band mixer		-8	
64-Input summer		17	
Signal at output of 64-input summer	-75.7		19.34
Wide-band i-f amplifier		58.3	
Up-converter		-10	
Band-pass filter		-2	
Traveling-wave tube (TWT) amplifier		+40	
Signal at TWT output	10.6		19.34
64-way power divider		-20	
High-level mixer		-11.5	

Table 3-1. Summary of signal and noise levels.

Parameter	Signal Levels (dbw)	Component Gain (db)	Signal-to Noise Ratios(db)
Signal output of high-level mixer	-20.9		17.0
Transmitting diplexer		-1.8	
Signal level at transmitting element	-22.7		
Transmitting element gain		11.6	
Number of elements squared $(64)^2$		36.1	
Effective radiated power (N^2gp)	25		19.3
<u>Pilot Signals</u>			
Received signal level (isotropic)	-125.5		
Receiving antenna element gain		11.6	
High-pass filter		-1.8	
First mixer		-8	
Wide-band i-f preamplifier		30	
Signal at preamplifier output	-93.7		
Triplexer		-1	
First pilot mixer		14	
12-M Hz pilot i-f amplifier		40	
Quadraplexer		-2.5	
Signal at quadraplexer output	-43.2		21.34
Second pilot mixer		-3.5	
Receiving pilot i-f amplifier		19.7	
Transmitting pilot i-f amplifier		40	
Signal at receiving pilot i-f output	-27		21.34
Signal at transmitting pilot i-f output	-6.7		21.34

Table 3-1 (continued). Summary of signal and noise levels.

4.0 ANTENNA ELEMENT AND ARRAY DESIGN

4.1 ELEMENT CONSIDERATIONS

The array initially considered for this system was an 8-by-8 planar array with uniform interelement spacing. For this type of an array, with each element connected to an independent matched generator or load, the total gain in the beam-pointing direction is N times the effective gain, g, of a typical element. Where g is measured in the presence of all other elements terminated in matched loads and the number of elements is N and is assumed to be sufficiently large so that most elements see similar environments.

Each element radiates 1/N of the total power P. Therefore, the available power per element, p, is

$$p = \frac{P_a}{N} \tag{4-1}$$

and the effective radiated power (ERP) is

$$\mathrm{ERP} = P_a G = N^2 gp \tag{4-2}$$

The required coverage region is a cone of half angle 15 degrees. The element factor must cover this region. To suppress grating lobes for the scanned beams, the element factor must also be small outside this region. It was therefore assumed that the element factor 3 db points fall at ±15 degrees. If the element pattern is assumed to be essentially that of a uniformly illuminated segment of the aperture, its pattern has the form

$$\frac{\sin\left(\frac{kL}{2}\sin\theta\right)}{\frac{kL}{2}\sin\theta}$$

where k is the free space propagation constant, L is the edge length of the element area, and θ is the angle off broadside in the principal plane. At $\theta = \theta_o = 15^o$, this pattern must be down 3 db from its maximum value. This requires that L/λ be 1.714. The area gain is then approximately

$$g_a = 4\pi\left(\frac{L}{\lambda}\right)^2 = 36.9 \tag{4-3}$$

To account for losses due to spacing, coupling, and other factors, it was assumed that actual element gain is 1 db less so that g = 29.2 at broadside and is half that value at 15 degrees from broadside in the principal plane, gain = 11.6 db.

The gain was given by

$$G_{max} = (64)\ (29.2) \text{ or } 32.8 \text{ db}$$

$$G_{min} = (64)\ (14.6) \text{ or } 29.8 \text{ db} \tag{4-4}$$

These values for element and array gain were used during the system design.

Because of the periodic structure of an 8-by-8 planar array and the large interelement spacing required to realize the desired gain, grating lobes exist. The element factor must fall off sharply, about 3 db, at the edge of the coverage angle, to suppress these lobes. To eliminate the gating lobes, it would be desirable to have the elements arranged in a nonuniform fashion which would avoid the periodic structure inherent in an 8-by-8 array. If the grating lobes can be avoided in the array factor, the element factor may be broadened somewhat which will reduce the peak gain of the array but will not decrease the minimum gain for the beam scanned ±15 degrees from broadside.

A detailed consideration was given to the preliminary design of a 64-element, circularly polarized antenna which is capable of scanning a 30-degree cone. To achieve high gain for this antenna, the radiation pattern of each element should be sufficiently directive, and consequently the separation between the elements must be large enough, so that the effective aperture areas of the elements do not overlap. The directivity of each element is limited by the scanning requirement. The ideal element factor is of the form

$$E(\theta, \phi) = \begin{cases} \sec\theta & 0^\circ \leq \theta \leq 15^\circ \\ 0 & 15^\circ < \theta \leq 90^\circ \end{cases}$$

that (1) the antenna gain remains essentially constant as the beam scans off broadside and (2) the power radiated in the undesired regions is minimized.

Of the various array elements, such as horns, array of cross-slots, and helices, helices seemed to be the most suitable as the basic radiating elements. Because of the large separation between elements and the desire for no grating lobes, the usual uniform spacing of the elements must be avoided. A promising approach is to arrange the elements in concentric circles (Figure 4-1) with every element approximately the same distance from its neighbors. The average separation between elements in this circular array is about 2-1/4 wavelengths.

Calculations were made for the radiation patterns of the circular array shown in Figure 4-1 for various scan angles - with and without the assumed element factor (Figure 4-2). The results of those calculations are presented in Figure 4-3. The radiation pattern of an 8-by-8 uniformly spaced (2-1/4 wavelengths) array is given in Figure 4-4 for comparison.

4.2 ANTENNA GAIN STATISTICS

Because of errors in phase and amplitude for each of the 64 elements, the average antenna gain will be reduced, and random deviations can be expected. The first step in analyzing

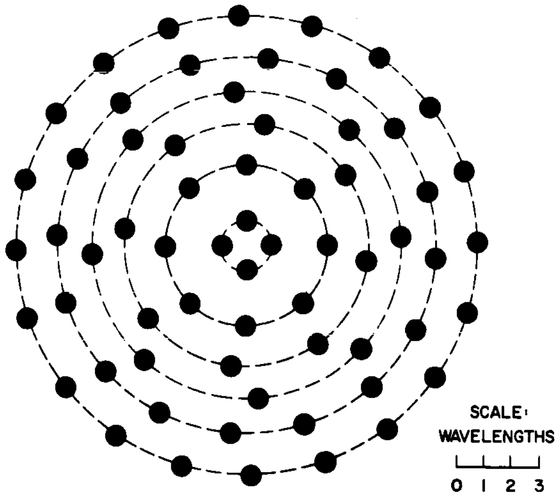


Figure 4-1. Arrangement of elements of 64-element circularly polarized array.

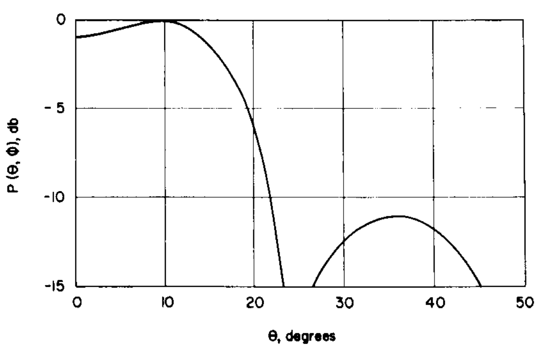
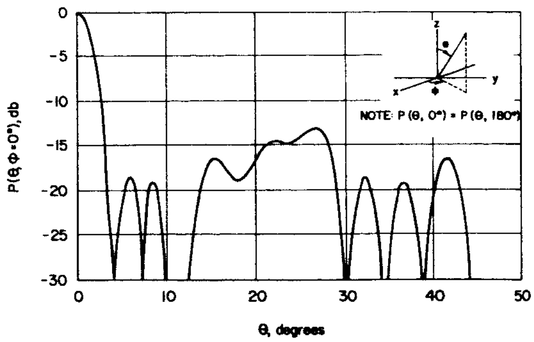
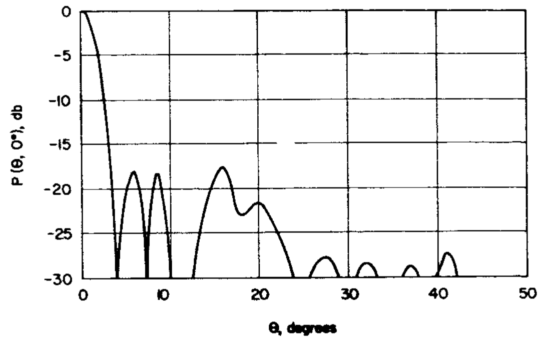


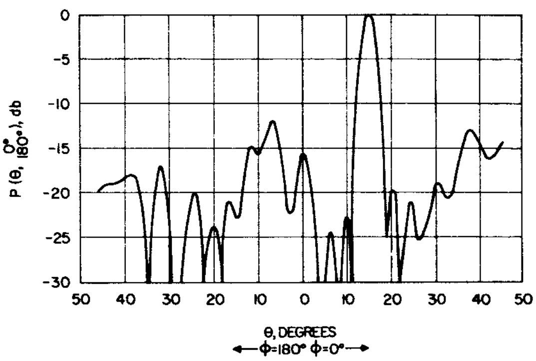
Figure 4-2. Assumed element factor of circular array.



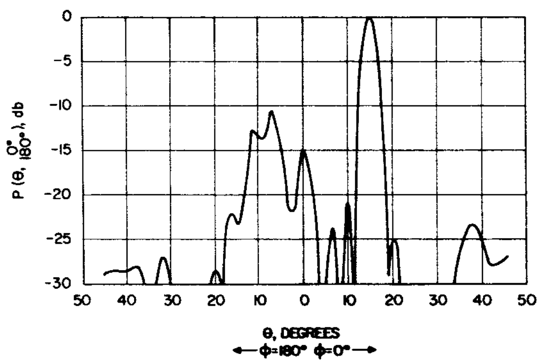
Without element factor



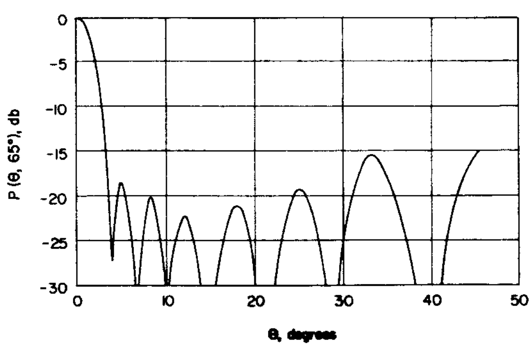
With element factor



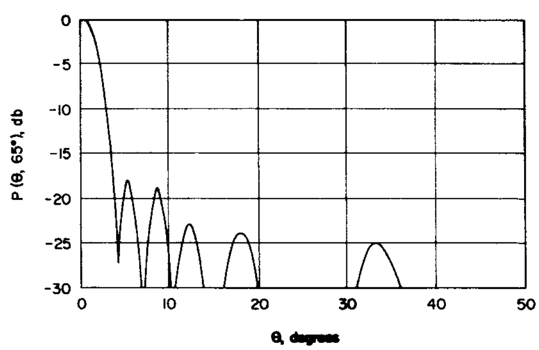
Without element factor



With element factor



Without element factor



With element factor

Figure 4-3. Calculated radiation patterns for circular array for various scan angles with and without element factor.

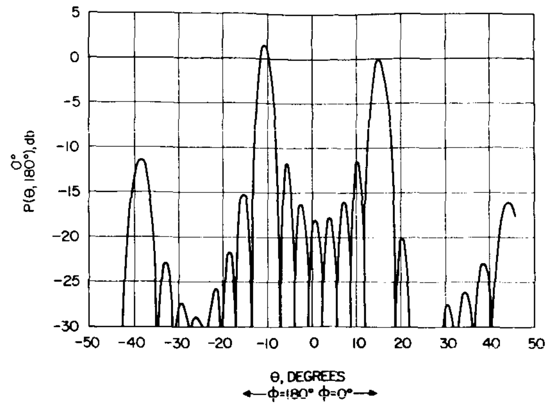


Figure 4-4. Radiation pattern of 8-by-8 planar array at a 15-degree scan angle with element factor included and an interelement spacing of 2.25λ.

such an effect is to calculate the statistics of antenna gain assuming independent phase and amplitude errors at each element. (The independent amplitude error assumption is poor because of the usual definition of antenna gain. However, the difference will be discussed later.) For this model the gain will be defined by the ratio of the power density at broadside in the actual array to the power density at broadside in the error-free array.

From a derivation stemming from probability theory* the average power gain of an error-prone antenna can be found to be

*J. E. Howard, "System Performance," Section 5, Traveling-Wave Phase Shifter-Phase II, (Contract (AF 30(62)-3532), F.A. Terrio and W.H. Kummer, eds.; Hughes Aircraft Company, Report No. P66-62, Culver City, California, February 1966.

$$\overline{G} = \frac{\overline{P}(\theta_o, \phi_o)}{{}_oP(\theta_o, \phi_o)} = e^{-\sigma_\delta^2} \left[1 + \frac{(\sigma_\delta^2 + \sigma_\epsilon^2) \sum_{i=-N}^{N} ({}_of_i^2)}{\left(\sum_{i=N}^{N} {}_of_i \right)^2} \right] \tag{4-5}$$

where

$\overline{P}(\theta_o, \phi_o)$ = expected power density at broadside

${}_oP(\theta_o, \phi_o)$ = the error-free power density at broadside

σ_δ = the rms phase error for the signals at the elements

σ_ϵ = the rms amplitude error for the signals at the elements

N = the number of elements

f_i = the excitation coefficient of the i^{th} element.

For a 64-element uniformly illuminated array, there results

$$\overline{G} \approx 1 - \sigma_\delta^2$$

$$\sigma_G \approx \frac{\left(256\ \sigma_\epsilon^2 + \sigma_\delta^4\right)^{1/2}}{64} \tag{4-6}$$

where σ_G is the rms error in gain. It is now obvious that, unless $\sigma_\epsilon^2 \ll \sigma_\delta^4$,

$$\sigma_G \approx \frac{1}{4} \sigma_\epsilon \tag{4-7}$$

The probability distribution of the random gain variability may be approximated by a χ^2 distribution. However for a high number ($\nu > 30$) degrees of freedom, the χ^2 distribution approaches a normal distribution. And, at any rate, the normal distribution percentage points will yield more conservative answers to the gain statistics questions.

The gain statistics are described by

$$\text{Prob}\left[\text{Gain} > \overline{G} - n\sigma_G\right] = P_n \tag{4-8}$$

For the normal distribution,

$P_{1.0} = 84.1$ percent

$P_{2.0} = 97.7$ percent

$P_{2.3} = 99.0$ percent

$P_{2.6} = 99.5$ percent

where the subscript indicates the multiplier for the standard deviation.

For determination of the gain and phase tolerances to be imposed upon the electronics for this system, it is desirable to have a gain degradation as small as is reasonable with a high confidence level. The gain degradation which will be allowed for this system is less than 1/2 db with a 99 percent confidence.

The 99 percent gain level is given by

$$1 - \sigma_\delta^2 - 2.3\left(\frac{\sigma_\epsilon}{4}\right)$$

For a total gain degradation less than 1/2 db from the no-error gain (with 99 percent confidence) there is required

$$\frac{1}{2}\text{ db} > 10 \log_{10} (1 - \sigma_\delta^2 - 0.575\,\sigma_\epsilon) \tag{4-9}$$

This requirement means that any of the following combinations of rms errors would be permissible:

$\sigma_\delta < 5^\circ$, $\sigma_\epsilon = 0.192$ (1.7 db)

$\sigma_\delta = 10^\circ$, $\sigma_\epsilon = 0.139$ (1.2 db)

$\sigma_\delta = 15^\circ$, $\sigma_\epsilon = 0.0695$ (0.6 db)

$\sigma_\delta = 19^\circ$, $\sigma_\epsilon = 0$

These results assume that the main beam peak levels for independent errors directly describe antenna gain. However, as used here, true antenna gain must be defined in terms of a constant input power to the antenna. So, in terms of this definition, the main beam peak levels will describe the antenna gain when errors are assumed that are correlated so that total power input is constant. This special type of correlation restricts the antenna gain variations so that larger errors than those quoted previously are permissible. In view of these considerations, the stated phase and amplitude error specification for .he system electronics seem conservative.

5.0 ANTENNA SYSTEM ELECTRONICS

The components of the antenna system may be classified into two distinct categories: the microwave components which operate in the frequency range of 7 to 8 GHz and the i-f circuitry that operates at frequencies below 800 MHz.

All the microwave components appear to be technically feasible as presently specified in this report. State-of-the art factors in the microwave area relate to the phase-tracking requirements. The input filter which has conflicting requirements of low loss, sharp attenuation, and small size may also be near the limit of present designs.

Since all the i-f circuitry will be designed in-house, the specifications of the various blocks are not presented in as much detail as those for the microwave components. Much of the study effort relating to the i-f has been devoted to ascertaining the feasibility and possible configuration of the wide-band preamplifiers and the triplexer filter for channel splitting. The wide-band preamplifier utilizing high-frequency transistors at a reasonable cost with the noise figure specified will push the state-of-the-art.

The descriptions and specifications of components are made with reference to the block diagram of the integrated system shown in Figure 5-1. Component numbers refer to corresponding block numbers on the diagram.

5.1 HIGH-PASS FILTER (NUMBER 1)

5.1.1

The function of the input high-pass filter (see block No. 1 on Figure 5-1) is to provide rejection of the transmitted signals and prevent interference with the receiver. These signals range from 7.125 to 7.475 GHz. The filter pass-band of 7.825 to 8.175 GHz must encompass the two received information channels, of 125 MHz bandwidth each, and the four pilot signals received from the various ground stations. There is a separate filter connected to each of the 64 antenna elements, and each filter provides an output to one of the 64 low-level first mixers.

5.1.2

Design considerations include size and package configuration that will allow direct mounting and connection to the antenna element and short connection length to the mixer. Trade-off factors are low insertion loss versus high rejection of the transmitted signal. A difficulty in determining an optimum transmitter rejection is due to the presently unknown mutual coupling factor between transmitting and receiving antenna elements. To be safe, a 60 db rejection factor has been established. A mixer dynamic range also needs to be established more carefully. Further study will be necessary to determine if the rejection characteristic could be relaxed to reduce the insertion loss factor which increases receiver performance.

Strip transmission line construction, which is less expensive and lighter in weight, was ruled out by its high insertion loss which would be on the order of 4 to 5 db. Waveguide construction gives the desired electrical characteristics, but the size and weight are excessive. Interdigital and coaxial cylinder designs appear to meet the space requirements. Reasonable specifications from a system aspect can be established for an interdigital filter. A 5 or 6-section filter is required which results in a somewhat undesirable insertion loss factor of 1.8 db. A coaxial design has been offered that appears to provide better insertion loss with only a small increase in size and weight. However, technical details of this design have not been evaluated; thus there is insufficient confidence in the device to revise the specifications at this time.

5.1.3 Specifications

5.1.3.1 Pass Band: 7.825 to 8.175 GC; 350 MHz or greater bandwidth.

5.1.3.2 Insertion Loss in Pass Band: 1.8 db maximum required, 0.5 db desired.

5.1.3.3 Phase Shift in Pass Band: minimized.

5.1.3.4 Reject Band: 7.125 to 7.475 GC.

5.1.3.5 Attenuation in reject band: 60 db minimum.

5.1.3.6 Phase Tracking Error between units: ±5 degrees maximum.

5.2 FIRST MIXER (NUMBER 2)

5.2.1

One input mixer (No. 2 in Figure 5-1) is used for each receiving antenna element. The complete received signal from the high-pass filter is converted to an i-f signal with a frequency range of 450 to 800 MHz. A local oscillator frequency of 8.625 GHz is used. The mixers must provide good relative phase tracking, low conversion loss, and low noise.

5.2.2

The mixer will be a diode balanced mixer providing a single-ended i-f output. At present, both cavity mixers and strip transmission line mixers are being considered. The latter design, offered by a microwave component vendor, appears to have too high a conversion loss to be feasible; however, the technique offers the attractive possibility of a design integrated with a strip transmission line local oscillator power divider. This approach is being explored at the present time.

Problems involving the mixers include phase tracking, proper termination of the mixer ports to suppress the image signal generated within the mixer, and the possibility of excessively high conversion loss resulting from the wide-band and high i-f output requirements.

5.2.3 Specifications

5.2.3.1 R-F Signal Input: 7.825 to 8.175 GHz, -125.5 dbw pilot and -115.5 dbw information signals.

5.2.3.2 Local Oscillator Input: 8.625 GHz, 4 mw nominal, 5 mw maximum required.

5.2.3.3 Mixer Output to i-f amplifier: 450 to 800 MHz.

5.2.3.4 VSWR at input: $<1.6:1$.

5.2.3.5 Phase Tracking Error between units: $\pm 4^{\circ}$ maximum.

5.2.3.6 Conversion Loss, L_c, measured from r-f input to i-f output of mixer: $L_c = 8$ db maximum, L_c variation between units ± 1 db.

5.2.3.7 Noise Figure: 9 db maximum, based on i-f amplifier noise figure of 1.5 db (single side-band noise figure).

5.3 64-WAY POWER DIVIDER (NUMBER 3)

5.3.1

The local oscillator signals for the first mixers are supplied from the outputs of a 64-way power divider (No. 3). The power divider splits the energy from a 8.625 GHz local oscillator (No. 4) equally in such a manner as to maintain equal phase shift of the output signals.

5.3.2

Several ways of implementing the power dividers are under consideration: (1) four separate 16-way power divider assemblies driven by a four-way power divider, (2) a single 64-way design, or (3) a combined mixer and power divider design. Assembly size and configuration, phase tracking error between outputs, power division error, and isolation between outputs are important. The use of a strip transmission line power divider combined with a hybrid output circuit for isolation between outputs is presently planned. The power divider will be mounted in the electronic processing assembly close to the mixers.

5.3.3 Specifications

5.3.3.1 Operating Frequency: 8.625 GHz.

5.3.3.2 Insertion Loss: 20 db maximum, measured between input and each output separately (If the device were lossless, a power ratio of 18 db would be present).

5.3.3.3 Power Level Input: -4 dbw nominal.

5.3.3.4 Power Division Error: ±0.5 db maximum.

5.3.3.5 Relative Phase Tracking: $\pm 2^{\circ}$ maximum, measured between outputs.

5.3.3.6 Isolation between outputs: >20 db.

5.4 MICROWAVE LOCAL OSCILLATORS (NUMBER 4)

5.4.1

Two microwave local oscillators are used in the system. The first local oscillator, at a frequency of 8.625 GHz, provides the reference signal to convert the received signal to i-f for amplification and processing. The up-converter local oscillator, at 7.507 GHz, provides the reference for conversion of the processed i-f signal back up to microwave frequency for further amplification, processing, and transmission.

5.4.2

To save on cost in the present prototype, solid-state local oscillator units are not provided as integral parts of the system. These components are readily available and the substitution of less expensive oscillators will not alter the performance of the system. Mechanical layouts include the allocation of space for solid-state local oscillators which could be added at a later date.

The type of oscillator under present consideration is a klystron to be stabilized by the use of an oscillator synchronizer such as the Dymec 2650A. Oscillator tubes, synchronizers, and power supplies are to be rack-mounted remotely located from the electronic processing assembly.

5.4.3 Specifications — First Local Oscillator

5.4.3.1 Operating Frequency: 8.625 GHz.

5.4.3.2 Output Power: 400 mw nominal.

5.4.3.3 Stability: Long term $1/10^6$ per week
Short term $1/10^8$ averaged over one second.

5.4.4 Specifications — Second Local Oscillator

5.4.4.1 Operating Frequency: 7.507 GHz.

5.4.4.2 Output Power: 100 mw nominal.

5.4.4.3 Stability: Long term $1/10^6$ per week
Short term $1/10^8$ averaged over one second.

5.5 WIDE-BAND PREAMPLIFIER AND TRIPLEXER (NUMBER 5)

5.5.1

Sixty-four preamplifiers and triplexers are used in the system to amplify and distribute the i-f output signals from the first mixers. Sufficient gain is provided to establish the receiver noise figure. A triplexer at the output of each preamplifier separates the received signal into three components: the two information channels and a pilot channel that includes all four pilots. The filtering for the information channels establishes the receiver bandwidth for each of these two channels.

5.5.2

The preamplifiers will be individually packaged and mounted integrally with the front end mixers (No. 2). It is expected that eight cascaded stages using T1XM101 transistors (or the equivalent) and the strip transmission line techniques will be the design approach. The problems involved will be difficulty of gain and phase tracking over an octave bandwidth.

The triplexers will be composed of a three-section low-pass filter, a two-pole Butterworth band-pass filter, and a three-section high-pass filter. They will be integrated with the pilot processing unit modules. The problem areas are the same as those of the preamplifiers.

5.5.3 Specifications—Preamplifiers

5.5.3.1 Bandwidth: 450 MHz - 800 MHz (1 db bandwidth)

5.5.3.2 Gain: 30 db ± 1 db over the pass-band

5.5.3.3 Phase Tracking: ±6^{o} over the 1-db pass-band

5.5.3.4 Noise Figure: 7 db maximum

5.5.3.5 Input Impedance = Output Impedance: 50Ω

5.5.3.6 VSWR < 1.5:1 at input and output

5.5.4 Specifications — Triplexers

5.5.4.1 Low-pass Filter

Insertion Loss: 1 db maximum over pass-band

Phase Tracking: ±5^{o} over pass-band

Rejection: 30 db at 622 MHz

Pass-band: 400 - 575 MHz

5.5.4.2 Band-pass Filter

Insertion Loss: 1 db maximum over pass-band

Phase Tracking: ±5^{o} over pass-band

Rejection: 60 db at 575 MHz and 675 MHz

Pass-band: 620 to 630 MHz

5.5.4.3 High-pass Filter

Insertion Loss: 1 db maximum over pass-band

Phase Tracking: ±5^{o} over pass-band

Pass-band: 675 to 800 MHz

5.6 PILOT PROCESSING (NUMBER 6)

The four pilot signals from the central band-pass filter of the triplexer (No. 5) are converted to a lower intermediate frequency in the first pilot mixer (No.6A) amplified in an i-f amplifier (No.6B) separated into individual signals by a quadruplexer (No. 6B) and up-converted (No. 6C) to an appropriate frequency to provide reference signals to the wide-band mixers and high-level mixers. (These components are all included as No. 6: Pilot Processing on Figure 5-1.)

5.6.1 First Pilot Frequency Conversion (No. 6A)

The first pilot frequency conversion circuits consist of a 613-MHz local oscillator, a 64-way power divider, and 64 mixer circuits. The output signal from the mixer consists of the four pilot signals at 9, 11, 13, and 15 MHz, respectively.

The oscillator will be composed of a fifth overtone crystal oscillator at 102.167 MHz, a buffer amplifier driving a transistor doubler, and a driver amplifier at 204.334 MHz driving a transistor tripler using a 2N4012 overlay device. The power divider will be constructed with quarter-wave coaxial transformers, utilizing a 4-way divider as a basic building block. The mixers will be common transistor mixers with a conversion gain.

These circuits will be operating at a discrete frequency and should not present any difficulties.

5.6.1.1 Specifications — Oscillator

Center Frequency: 613 MHz ±600 Hz

Output Power: + 25 dbm ±2 dbm

Spurious Noise: - 40 db

5.6.1.2 Specifications — 64-way Power Divider

Isolation: 30 db minimum at 613 MHz

Distribution Loss: 25 db maximum

Output Coherency: $\pm 2^{o}$ maximum spread

Output Impedance: 50Ω

Output Power: 0 dbm ±0.5 db

5.6.1.3 Specifications — Mixer

Signal Input: 622 MHz to 628 MHz at -67 dbm
Input impedance = 50 Ω

Local Oscillator Input: 613 MHz at 0 dbm

Input impedance = 50 Ω

Conversion Gain: 14 db ± 1 db

Signal Output: 8-12 MHz

Output impedance = 50Ω

5.6.2 12-MHz Pilot I-F Amplifier and Quadruplexer (No. 6B)

The composite signal of the four pilot signals from the first pilot mixer (No. 6A) is amplified and fed to a quadruplexer, in which the pilots are separated. The quadruplexer establishes the bandwidth of each of the four pilot channels and raises the pilot signal-to-noise ratio to a 21-db level.

The frequency of the i-f amplifier is somewhat flexible since the pilot signals are eventually up-converted in frequency. The factors affecting selection of the frequency are a sufficiently wide bandwidth to cover all the pilots and a low enough center frequency to facilitate the design of the narrow-band quadruplexer filters. The choice of center frequency of 12 MHz permits an 8 MHz i-f bandwidth and filters with 150 KHz noise bandwidths.

The i-f amplifier will consist of three wide-band transistor stages or a single monolithic video amplifier. The quadruplexer will be composed of four 2-pole Butterworth filters in parallel. The frequencies of the filters are sufficiently separated so that interaction at the input should not present a difficult problem in the design.

5.6.2.1 Specifications — I-F Amplifiers

Gain: 40 db ± 1 db

3-db Band-pass: 8 MHz to 16 MHz

Input Impedance: 50 Ω

Output Impedance: matched to Quadruplexer

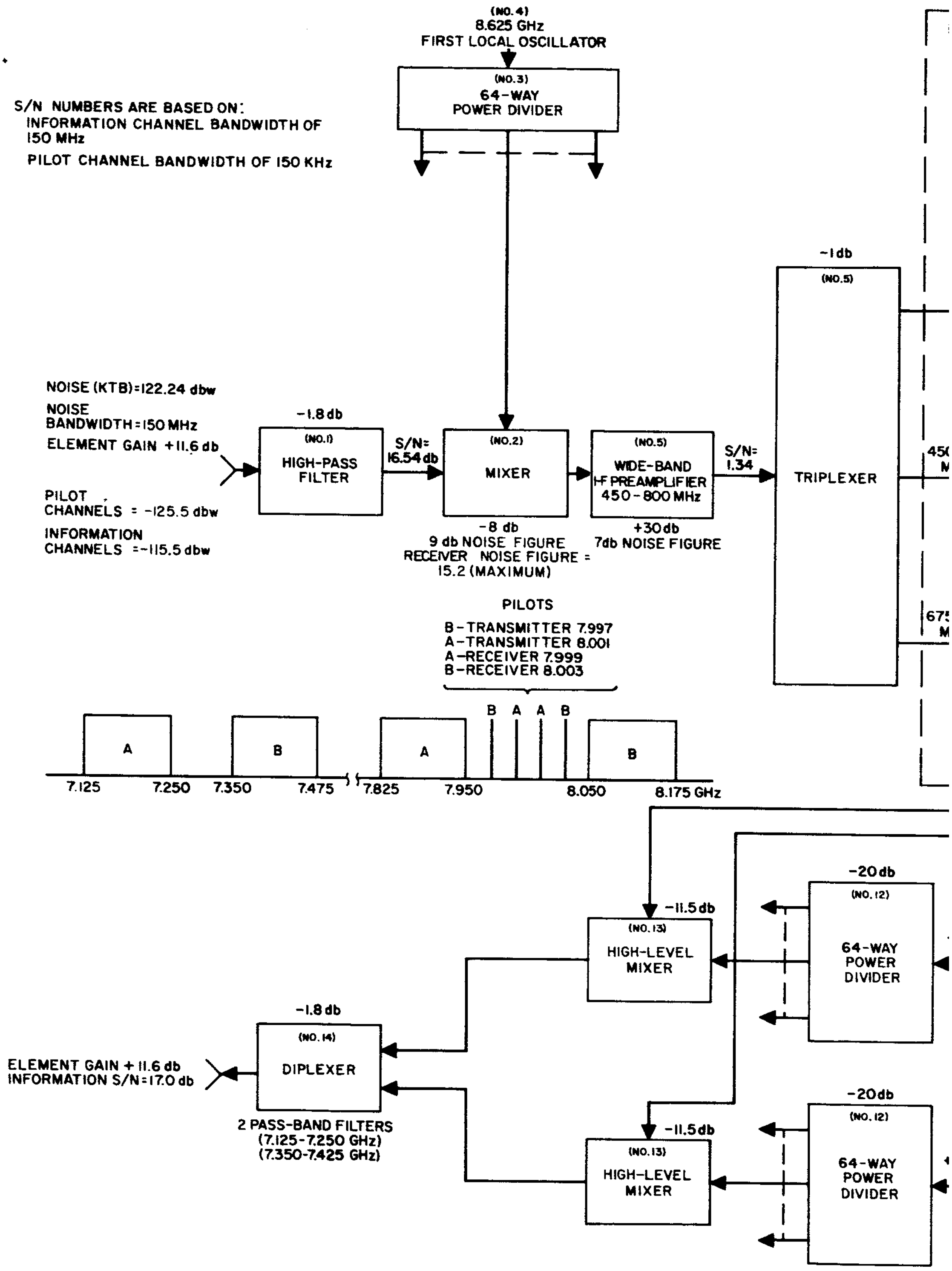
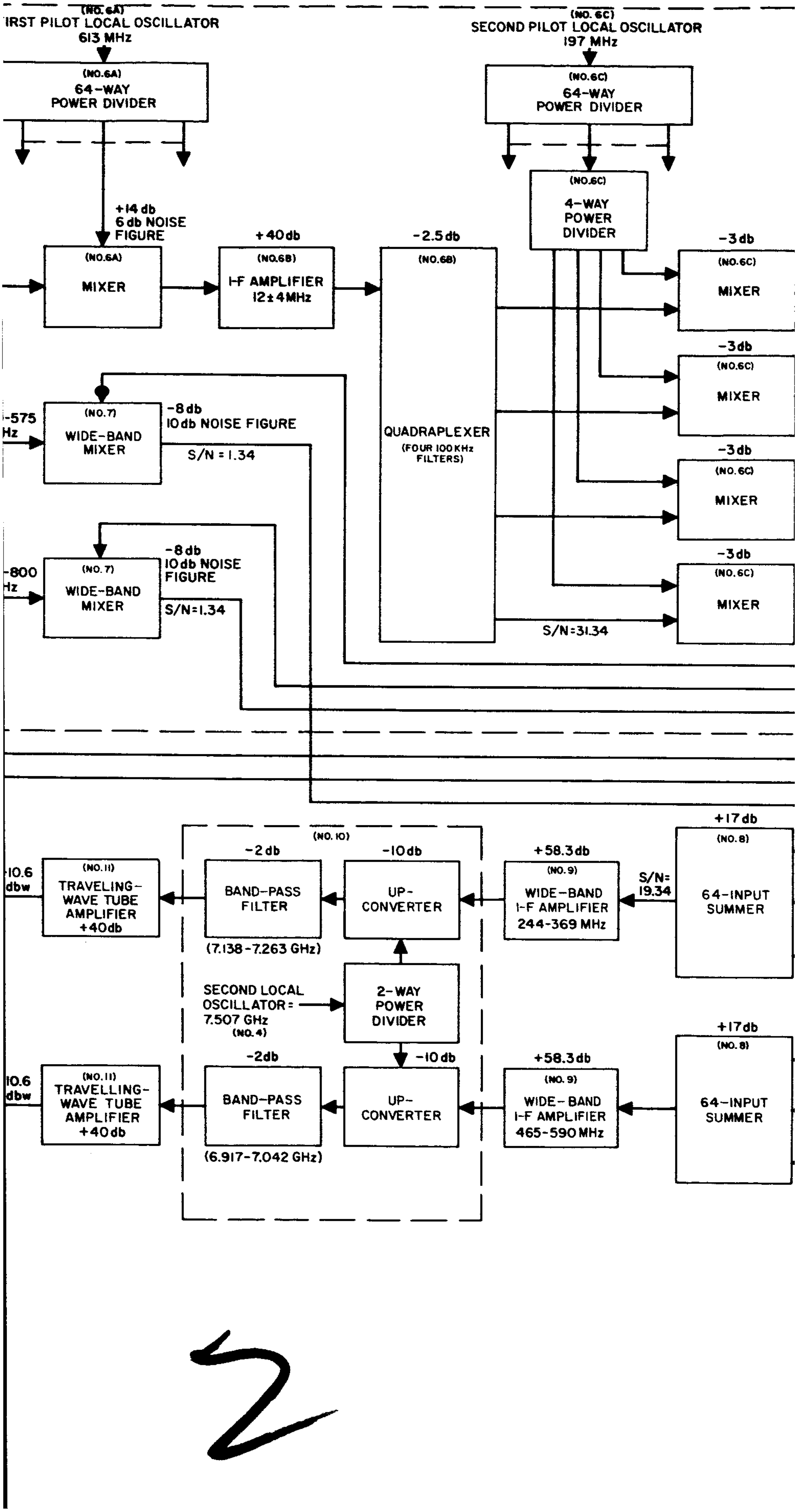
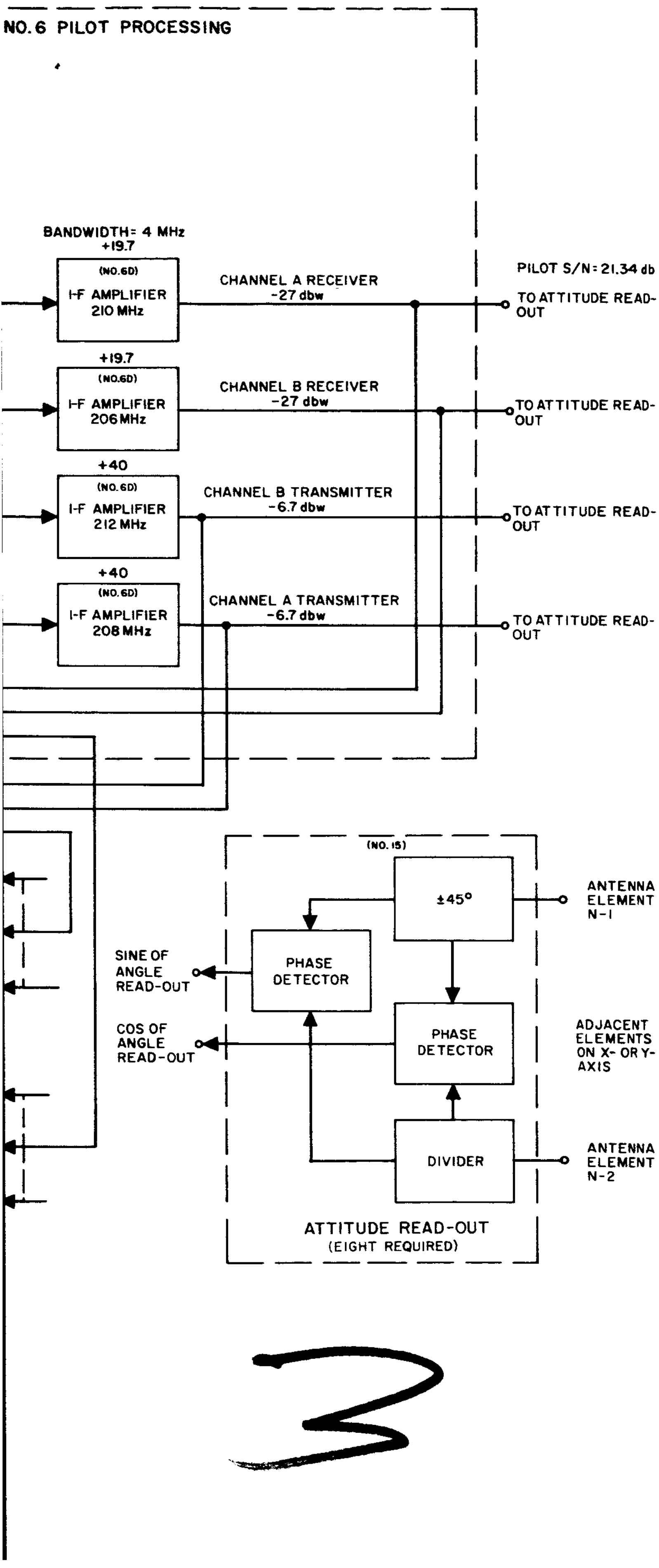


Figure 5-1. Integrated system block diagram.



(NO. 6A)
IRST PILOT LOCAL OSCILLATOR
613 MHz
(NO.6A)
64-WAY
POWER DIVIDER
+14 db
6 db NOISE
FIGURE
(NO.6A)
MIXER
+40 db
(NO.6B)
I-F AMPLIFIER
12±4 MHz
-2.5 db
(NO.6B)
QUADRAPLEXER
(FOUR 100 KHz FILTERS)
(NO. 6C)
SECOND PILOT LOCAL OSCILLATOR
197 MHz
(NO.6C)
64-WAY
POWER DIVIDER
(NO.6C)
4-WAY
POWER
DIVIDER
-3 db
(NO.6C)
MIXER
-3 db
(NO.6C)
MIXER
-3 db
(NO.6C)
MIXER
-3 db
(NO.6C)
MIXER
S/N=31.34
-575
Hz
(NO.7)
WIDE-BAND
MIXER
-8 db
10 db NOISE FIGURE
S/N = 1.34
-800
Hz
(NO. 7)
WIDE-BAND
MIXER
-8 db
10 db NOISE
FIGURE
S/N=1.34
(NO. 10)
10.6
dbw
(NO.11)
TRAVELING-
WAVE TUBE
AMPLIFIER
+40db
-2 db
BAND-PASS
FILTER
(7.138-7.263 GHz)
-10 db
UP-
CONVERTER
+58.3 db
(NO. 9)
WIDE-BAND
I-F AMPLIFIER
244-369 MHz
S/N=
19.34
+17 db
(NO. 8)
64-INPUT
SUMMER
SECOND LOCAL
OSCILLATOR =
7.507 GHz
(NO. 4)
2-WAY
POWER
DIVIDER
10.6
dbw
(NO.11)
TRAVELLING-
WAVE TUBE
AMPLIFIER
+40 db
-2db
BAND-PASS
FILTER
(6.917-7.042 GHz)
-10 db
UP-
CONVERTER
+58.3 db
(NO. 9)
WIDE-BAND
I-F AMPLIFIER
465-590 MHz
+17 db
(NO. 8)
64-INPUT
SUMMER
2


NO. 6 PILOT PROCESSING
BANDWIDTH = 4 MHz
+19.7
(NO. 6D)
I-F AMPLIFIER
210 MHz
CHANNEL A RECEIVER
-27 dbw
PILOT S/N = 21.34 db
TO ATTITUDE READ-OUT
+19.7
(NO. 6D)
I-F AMPLIFIER
206 MHz
CHANNEL B RECEIVER
-27 dbw
TO ATTITUDE READ-OUT
+40
(NO. 6D)
I-F AMPLIFIER
212 MHz
CHANNEL B TRANSMITTER
-6.7 dbw
TO ATTITUDE READ-OUT
+40
(NO. 6D)
I-F AMPLIFIER
208 MHz
CHANNEL A TRANSMITTER
-6.7 dbw
TO ATTITUDE READ-OUT
(NO. 15)
±45°
ANTENNA ELEMENT N-1
PHASE DETECTOR
SINE OF ANGLE READ-OUT
PHASE DETECTOR
COS OF ANGLE READ-OUT
ADJACENT ELEMENTS ON X- OR Y-AXIS
DIVIDER
ANTENNA ELEMENT N-2
ATTITUDE READ-OUT
(EIGHT REQUIRED)

3

5.6.2.2 Specifications — Quadruplexer

Center Frequency: 9 MHz, 11 MHz, 13 MHz, 15 MHz

Insertion Loss: -2.5 db at f_o

Rejection: 32 db at f_o ± 75 KHz

Input Impedance: matched to i-f amplifier

Output Impedance: 50 Ω

5.6.3 Second Pilot Frequency Conversion (No. 6C)

The separated pilot signals are converted up in frequency in the second pilot mixer to provide reference signals for phasing the input information signal and for phasing the output signal in each of the elements. The four mixers in each element are provided the local oscillator signal by means of a 4-way power divider located in the i-f module. The 4-way dividers receive their inputs from a 64-way power divider which in turn is driven by a 197 MHz source.

5.6.3.1 Specifications — Local Oscillator

Center Frequency: 197 MHz ± 197 Hz

Power Output: 32 dbm ± 2 dbm

Spurious Noise: -40 db

5.6.3.2 Specifications — 64-way Power Divider

Isolation: 30 db minimum at 197 MHz

Distribution Loss: 25 db maximum

Output Coherency: $\pm 2^{o}$ maximum spread

Output Impedance: 50 Ω

5.6.3.3 Specifications — Up-converters

Local Oscillator Input: 197 MHz at 0 dbm ± 0.5 db

Signal Input: 9, 11, 13, 15 MHz at -13 dbm

Conversion Loss: -3.5 db ± 0.5 db

Input Impedance: 50 Ω

Output Impedance: matched to second i-f inputs

Local Oscillator rejection: 30 db at 1-f output

Other sidebands: 30 db rejection at 1-f output

5.6.4 Pilot I-F Amplifiers (No. 6D)

The outputs of the second pilot mixers (No. 6C) are amplified in four separate pilot i-f amplifiers (No. 6D) to a level required by the wide-band mixers in the case of the receiving pilots and to a higher level needed by the high-level mixers in the case of the transmitting pilots.

Phase shifting circuitry will be incorporated into these amplifiers to align the phase shift of the individual elements so that the antenna system gain may be maximized.

Buffered outputs of the four pilot signals are available on external connectors for use as test points in adjustment of the interelement phase shifts. The outputs of two pairs of elements are used as reference signals in the attitude read-out circuits. However, the rest of the elements will retain the buffer circuits to provide test points.

The transmitting pilot amplifiers will consist of a limiting stage, a driver and phase shift stage, and a Class C power amplifier. The receiving pilot amplifiers will consist of a buffer and phase shift stage and a limiting amplifier stage. No problems are expected.

5.6.4.1 Specifications – Transmitting Pilot Amplifier

Input Impedance: As required

Output Impedance: 50 Ω

Output Power: 23.3 dbm ±0.2 db

Gain: 40 db ±0.2 db at center frequency

Selectivity: -25 db at f_o ±8.5 MHz

5.6.4.2 Specifications – Receiving Pilot Amplifier

Input Impedance: As required

Output Impedance: 50 Ω

Output Power: 3 dbm ±0.2 db

Gain: 19.7 ±0.2 db at center frequency

Selectivity: -25 db at f_o ±8.5 MHz

5.6.4.3 Specifications – Phase Circuitry

Phase shift ±45 degrees trim adjustment for both types of amplifiers

5.7 WIDE-BAND MIXERS (NUMBER 7)

5.7.1

The wide-band mixers (No. 7) receive the separated information signals from the high-pass and low-pass outputs of the triplexers (No. 5). The relative phases between elements are cancelled at this point by mixing with the receiving pilot signal. The mixer output in each information channel is now in phase with those of the corresponding mixers in the other elements. The 128 required mixers will use balanced diodes. Phase tracking and insertion loss are expected to be difficult problems. Special diodes such as hot carrier diodes may be required.

5.7.2 Specifications

5.7.2.1 Input Impedance: As required

5.7.2.2 Output Impedance: 50 Ω

5.7.2.3 Conversion Loss: -8 db

5.7.2.4 Local Oscillator Power: 3 dbm ±0.2 db

5.7.2.5 Signal: -55 dbm ±5 db

5.8 64-INPUT SUMMER (NUMBER 8)

5.8.1

The wide-band mixer outputs (No. 7) from the elements of each channel feed a 64-input summer to realize the receiving antenna gain and raise the signal-to-noise ratio by a factor of 18 db.

5.8.2

The design approach for the 64-input summers will be similar to that used with the i-f 64-way dividers; that is, summing networks to minimize insertion loss and provide 20-db isolation. Some difficulty is expected since the ports must track over a 125-MHz bandwidth in order to minimize delay distortion.

5.8.3 Specifications

5.8.2.1 Insertion Loss: 1 db maximum

5.8.2.2 Phase Tracking: $\pm 2^{\circ}$

5.8.2.3 Input Impedance: 50 Ω

5.8.2.4 Output Impedance: 50 Ω

5.9 I-F AMPLIFIER (NUMBER 9)

5.9.1

Two wide-band i-f strips are required. Their function is to amplify the respective information channel signals after the array gain has been realized.

5.9.2

The basic design of the wide-band i-f amplifiers will be similar to that of the preamplifiers; T1XM101 transistors (or the equivalent) can be used. It is expected that nine stages will be required for the 244-369 band and 14 stages for the 465-590 band. The high gain requirement will present some problems, but since each amplifier is a one of a kind item, the difficulty of repeatability with minimum parts is reduced.

5.9.3 Specifications

5.9.3.1 Pass-band: 244 to 369 MHz
465 to 590 MHz

5.9.3.2 Input Power: -45.7 dbm + 6.3 db - 3.7 db

5.9.3.3 Input Impedance: 50 Ω

5.9.3.4 Output Power: +13.7 dbm ± .5 db with AGC

5.9.3.5 Gain: 58.3 db ± 1 db over the pass-band

5.9.3.6 Output Impedance: 50 Ω

5.10 UP-CONVERTERS, BAND-PASS FILTER, TWO-WAY POWER DIVIDER (NUMBER 10)

5.10.1

After the Channel A information is amplified in the wide-band, 465 to 590 MHz, i-f amplifier, it is sent through an up-converter with an output of 6.917 to 7.042 GHz. Similarly, the Channel B information amplified in the wide-band, 244 to 369 MHz, i-f amplifier is sent through an up-converter with an output of 7.138 to 7.263 GHz. A single local oscillator at 7.507 GHz is used with a two-way power divider to supply both up-converters. The output of each up-converter is connected to a band-pass filter which passes the desired information band and rejects the image frequency band and local oscillator frequency.

5.10.2

Crystal diode up-converters are to be used. The local oscillator power required is estimated at -14.4 dbm which is 3 db higher than the input i-f power level. A local oscillator power level of 100 mw should be adequate.

The band-pass filters will be of interdigital or coaxial cylinder construction.

5.10.3 <u>Specifications – Up-converter, Channel A</u>

5.10.3.1 I-F Signal Input: 465 to 590 MHz, -17.4 dbw

5.10.3.2 Local Oscillator Input: 7.507 GHz, -14.4 dbw maximum required

5.10.3.3 Up-converter Output: 6.917 to 7.042 GHz

5.10.3.4 VSWR at input: $<1.6:1$

5.10.3.5 Insertion Loss measured from i-f input to r-f output: 10 db maximum

5.10.4 <u>Specifications – Up-converter, Channel B</u>

5.10.4.1 I-F Signal Input: 244 to 369 MHz, -17.4 dbw

5.10.4.2 Local Oscillator Input: 7.507 GHz, -14.4 dbw

5.10.4.3 Up-converter Output: 7.138 to 7.263 GHz

5.10.4.4 VSWR at input: $<1.6:1$

5.10.4.5 Insertion Loss measured from i-f input to r-f output: 10 db maximum

5.10.5 <u>Specifications — Band-pass Filter, Channel A</u>

5.10.5.1 Pass-band: 6.917 to 7.042 GHz

5.10.5.2 Insertion loss in pass-band: 2 db maximum

5.10.5.3 Rejection band: 7.442 to 8.097 GHz

5.10.5.4 Attenuation in rejection band: 30 db minimum

5.10.5.5 Attenuation at 6.517 GHz: 30 db minimum

5.10.6 <u>Specifications — Band-pass Filter, Channel B</u>

5.10.6.1 Pass-band: 7.138 to 7.263 GHz

5.10.6.2 Insertion Loss in pass-band: 2 db maximum

5.10.6.3 Rejection Band: 7.663 to 7.876 GHz

5.10.6.4 Attenuation in rejection band: 30 db minimum

5.10.6.5 Attenuation at 6.738 GHz: 30 db minimum

5.10.6.6 Attenuation at 7.507 GHz: 30 db minimum

5.10.7 <u>Specification — Two-way Power Divider</u>

5.10.7.1 Operating Frequency: 7.507 GHz

5.10.7.2 Insertion Loss: 3.5 db maximum, measured between input and each output separately

5.10.7.3 Power Level Input: 100 mw nominal

5.10.7.4 Power Division Error: ±0.3 db maximum

5.10.7.5 Relative Phase (measured between outputs): $\pm 2^{\circ}$ maximum

5.10.7.6 VSWR: 1.3 maximum

5.10.7.7 Isolation between outputs: 20 db minimum

5.11 TRAVELING-WAVE TUBE (NUMBER 11)

5.11.1

The function of each traveling-wave tube (No. 11) is to amplify an information channel after it has been up-converted and filtered. The output drives a 64-way power divider (No. 12).

5.11.2

The tubes are to be mounted in the electronic processing assembly with the high-voltage power supply mounted separately on the engineering model. Precautions are required in avoiding noise pickup in the high-voltage supply leads. Matching the output impedance of the tube is very important over the frequency range. There is a possibility that an isolator may be required in the output. The tubes are operated in their saturated region as only frequency modulated signals are used.

The TWT's used for the engineering model will be essentially off-the-shelf units instead of tubes especially designed for space applications. For this reason the efficiency will be quite low. However, the cost of high efficiency space tubes is much greater and the use of such a tube does not contribute to the demonstration of the system performance.

5.11.3 Specifications (Each Tube)

5.11.3.1 Operating Band: 6.917 to 7.263 GHz

5.11.3.2 Saturated Gain: 40 db minimum

5.11.3.3 Saturated Power Output: 11.5 watts (or 10.6 dbw) minimum assuming no loss from mismatch or isolator.

5.11.3.5 Power Supply: To include protective features

5.12 SIGNAL POWER DIVIDER (NUMBER 12)

5.12.1

The output of each traveling-wave tube (No. 11) is connected to a 64-way power divider (No. 12). Each information channel is split into the 64 elements again and fed to high-level mixers (No. 13).

5.12.2

The problem is similar to the 64-way power dividers required for the local oscillators. An added problem is that the range of frequencies involved is such that it may be difficult to use the same design for both channels.

5.12.3 Specifications – 64-Way Power Dividers

5.12.3.1 Operating Frequency: 6.917 to 7.263 GHz

or

Two Designs: Channel A, 6.917 to 7.042 GHz
Channel B, 7.138 to 7.263 GHz

5.12.3.2 Insertion Loss: 20-db maximum, measured between input and each output separately

5.12.3.3 Power Level Input: +10.6 dbw nominal

5.12.3.4 Power Division Error: ±0.5 db maximum

5.12.3.5 Relative Phase Tracking: $\pm 2^{\circ}$ maximum measured between outputs

5.12.3.6 Isolation between outputs: >20 db

5.13 HIGH-LEVEL MIXERS (NUMBER 13)

5.13.1

Sixty-four high-level mixers are used for each information transmit channel. The proper phase information is applied to the signal by mixing it with the transmitting pilot signals. The output of each mixer is applied to a diplexer input.

5.13.2

Design considerations include supplying about equal or greater local oscillator power than the signal to the crystal diode mixer which has an r-f input of -9.4 dbw. The physical size and arrangement has to be compatible with the sixty-four way power dividers and diplexers. A total of 192 connections is required.

5.13.3 Specifications — Channel A

5.13.3.1 R-F Input Range: 6.917 to 7.042 GHz, -9.4 dbw

5.13.3.2 Local Oscillator Input: 208 MHz, -6.7 dbw maximum

5.13.3.3 Output: 7.125 to 7.250 GHz

5.13.3.4 VSWR at input: <1.6:1

5.13.3.5 Conversion Loss: 11.5 db measured from r-f input to r-f output

5.13.4 Specifications — Channel B

5.13.4.1 R-F Input Range: 7.138 to 7.263 GHz, -9.4 dbw

5.13.4.2 Local Oscillator Input: 212 MHz, -6.7 dbw maximum

5.13.4.3 Output: 7.350 to 7.475 GHz

5.13.4.4 VSWR at input: <1.6:1

5.13.4.5 Insertion Loss: 11.5 db measured from r-f input to r-f output

5.14 TRANSMITTING DIPLEXER (NUMBER 14)

5.14.1

Sixty-four diplexers are used to combine the two transmitting channels, 7.125 to 7.250 GHz and 7.350 to 7.475 GHz, for coupling to the sixty-four transmit antenna elements.

5.14.2

The diplexer problem is similar to the input filter. Each unit is to be mounted for direct connection to an antenna element. Manufacturers consulted indicate that it is possible to meet 1.8 db maximum insertion loss in an interdigital design and between 1 and 1.5 db maximum insertion loss in a coaxial design.

5.14.3 Specifications

5.14.3.1 Pass-band Channel A: 7.125 to 7.250 GHz
Channel B: 7.350 to 7.475 GHz

5.14.3.2 Insertion Loss in pass bands: 1.8 db maximum

5.14.3.3 Power Level: 10 mw

5.14.3.4 Cross-over frequency: 7.300 GHz, 60 db minimum isolation between channels

5.15 ATTITUDE READ-OUT (NUMBER 15)

5.15.1

Two attitude read-out circuits (No. 15) are required for each transmitting and receiving pilot. One read-out angle with respect to the x-axis and one with respect to the y-axis.

5.15.2

The phase of two adjacent elements on the x-axis will be compared in quadrature phase detectors necessary to determine angle and sign. Eight of these circuits will be required with no expected problem areas.

5.15.2 Specifications

Input Impedance: as required

Output Impedance: 10 K

Input Power: 0 dbm ±0.1 db

5.16 POWER REQUIREMENTS

Power requirements were estimated for each sub-unit with the exception of the microwave local oscillators and are presented in Table 5-1.

Unit	Power (watts)	Quantity	Total Power (watts)
Wide-band i-f preamplifier	0.4	64	25.6
First pilot local oscillator	0.6	1	0.6
First pilot mixer	0.07	64	4.5
Second pilot i-f amplifier	0.2	64	12.8
Second pilot local oscillator	3.3	1	3.3
Second pilot mixer (up-converter)	0.07	256	17.9
Third pilot i-f amplifier			
Transmitting pilot	0.4	128	51.2
Receiving pilot	0.14	128	17.9
Wide-band i-f amplifiers	0.6	2	1.2
Attitude Read-out			
Transmitting pilots	0		
Receiving pilots	0.14	4	0.6
Traveling-wave tube	100.0	2	200
		Total	335.6

Table 5-1. Power requirements.

6.0 ELECTRONICS PACKAGING AND STRUCTURE

The system chosen for the engineering model allows for discrete functional devices that are realizable in the short development time allowed. The system is separated into three basic units — i-f electronics, transmitting antenna, and receiving antenna — which allows for variations in the overall shape factor and for possible reduction in weight. This flexibility is necessary since final satellite shape factors are not known. The form factors of the three basic units can also be easily changed to allow for special space allotments if required. The system weight at the present is 175 pounds which is considered a maximum with a design goal of 110 pounds, exclusive of the power supplies and solid-state X-band local oscillators.

6.1 LAYOUTS FOR PACKAGING

Detailed preliminary layouts of the integrated engineering system configuration were made to determine the feasibility of packaging it in a workable unit and to establish more accurate weight and volume estimates. Sketches of the layouts are shown in Figures 6-1 through 6-5. The size of the system is about 2 by 4 by 3 feet. The boxes housing the i-f electronics may be distributed differently than shown in the layouts to vary the shape of the system as desired.

The alternative dual-channel system configuration is shown in a side view in Figure 6-6 for comparison.

6.2 WEIGHT ESTIMATES

The estimated weight of the engineering model is 175 pounds. This figure does not include the weight of microwave local oscillators, the traveling-wave tube power supplies, or the system power supplies. The alternative configuration that was considered in this study would have weighed an estimated 250 pounds (see Section 2.1.1).

The weight breakdown of the various units of the integrated system is shown in Table 6-1, with the differential weights of the alternative system detailed in Table 6-2.

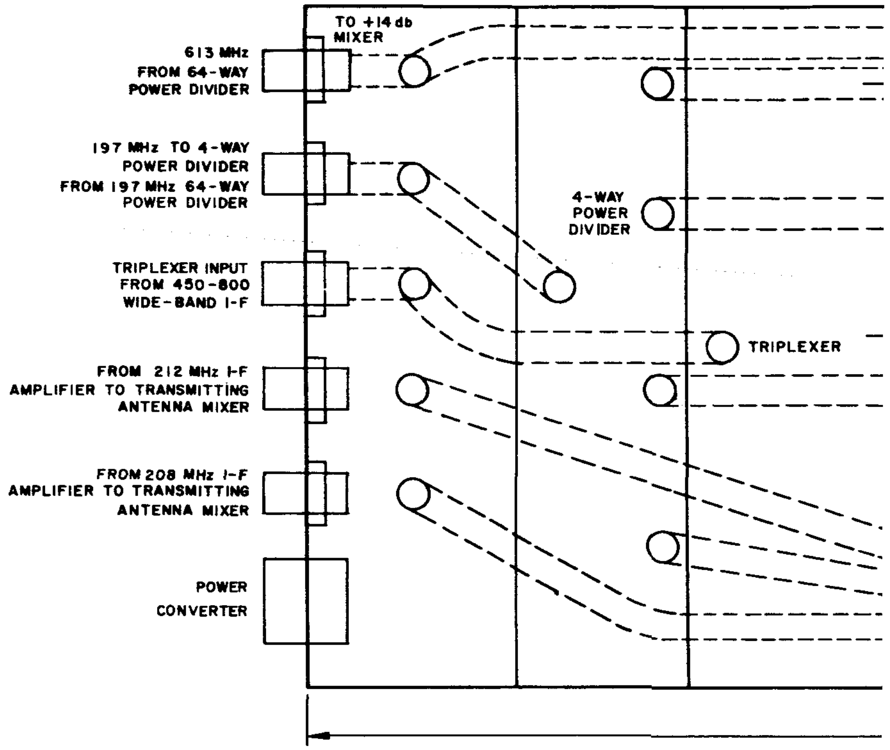
6.3 CONSTRUCTION TECHNIQUES

6.3.1 I-f Module

In construction of the 32 i-f modules required for the engineering model electronics unit (Figure 6-1), basic printed circuit boards will be used. The boards will contain all the i-f circuits except the wide-band i-f preamplifiers and the non-repetitive circuits. All other parts will be mounted on the boards which will then be dip-soldered. The egg crate for shielding, which consists of 0.010 copper-flashed aluminum, will then be affixed to the printed boards. Interconnections between functional circuits will be made with coaxial cables.

Two completed, basic, printed circuit boards will be jigged with a 1/32-inch copper-clad epoxy board separating them and foamed in place. The entire board assembly will be inserted into a copper- or silver-flashed aluminum investment casting and soldered in place. The remaining work will be to insert the coaxial connectors and the power connectors into the casting, make the appropriate connections, and attach the lids in the receiving slots in the casting.

The traveling-wave tubes will be mounted on finned aluminum blocks with a fan circulating air. Their location would vary in a final flight model. Layout of the main element rack containing



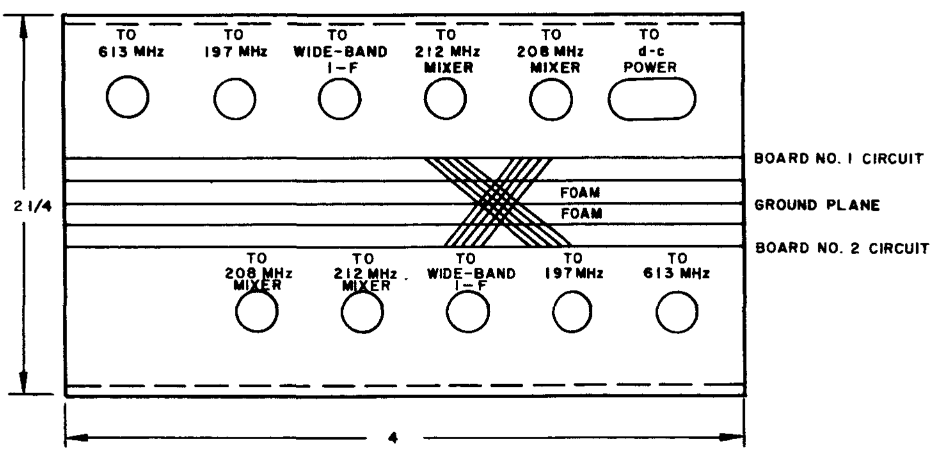
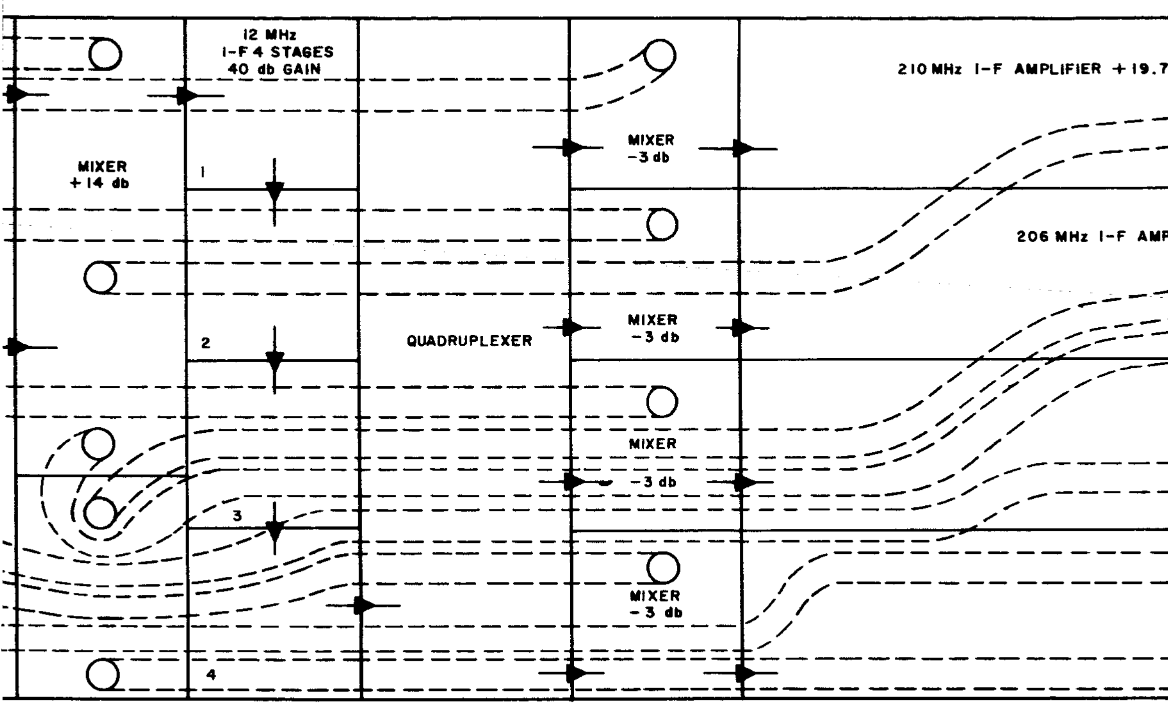


Figure 6-1. I-f module layout for integra
engineering model.



(DIMENSIONS ARE IN INCHES)

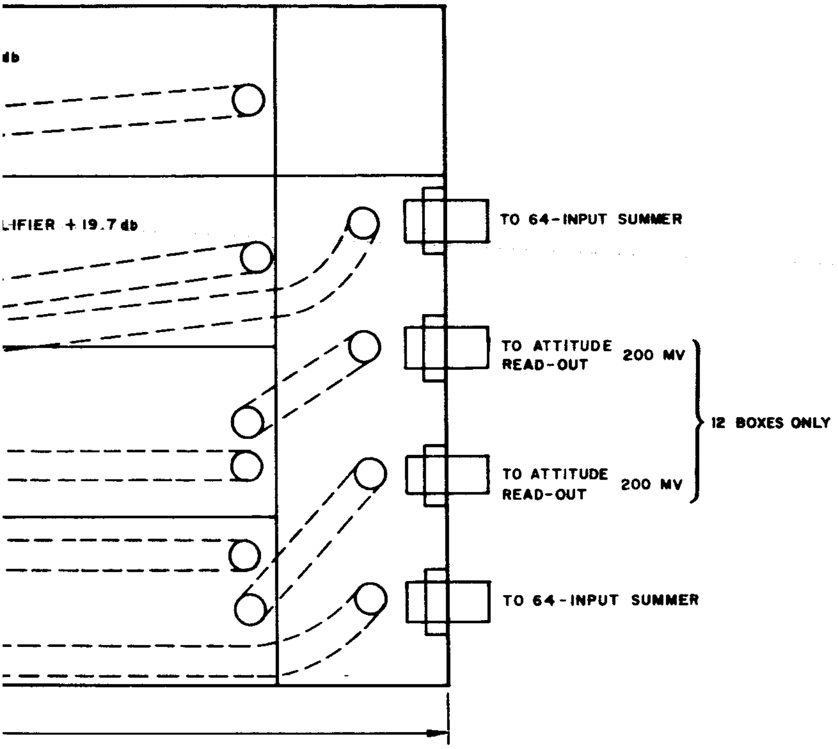
FOAM TO BE 3/16 THICK

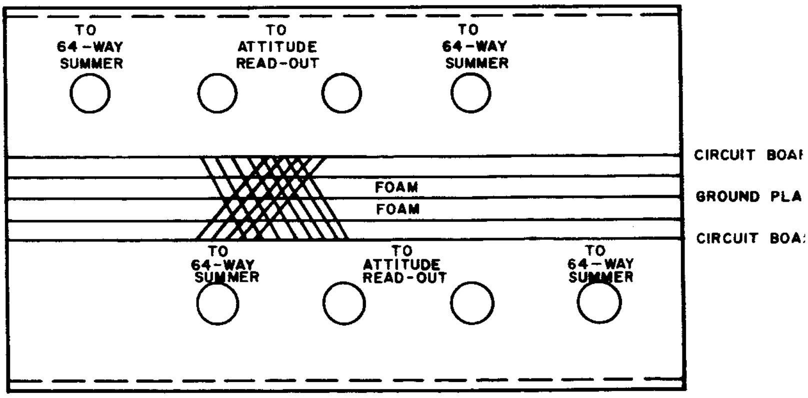
CENTER BOARD TO BE
(2) 1/32 BOARDS
BACK TO BACK

TOP CIRCUIT BOARD 1/32

2

ed system


db
LIFIER + 19.7 db
TO 64-INPUT SUMMER
TO ATTITUDE READ-OUT
200 MV
12 BOXES ONLY
TO ATTITUDE READ-OUT
200 MV
TO 64-INPUT SUMMER


TO 64-WAY SUMMER
TO ATTITUDE READ-OUT
TO 64-WAY SUMMER
CIRCUIT BOA
FOAM
GROUND PLA
FOAM
CIRCUIT BOA
TO 64-WAY SUMMER
TO ATTITUDE READ-OUT
TO 64-WAY SUMMER

3

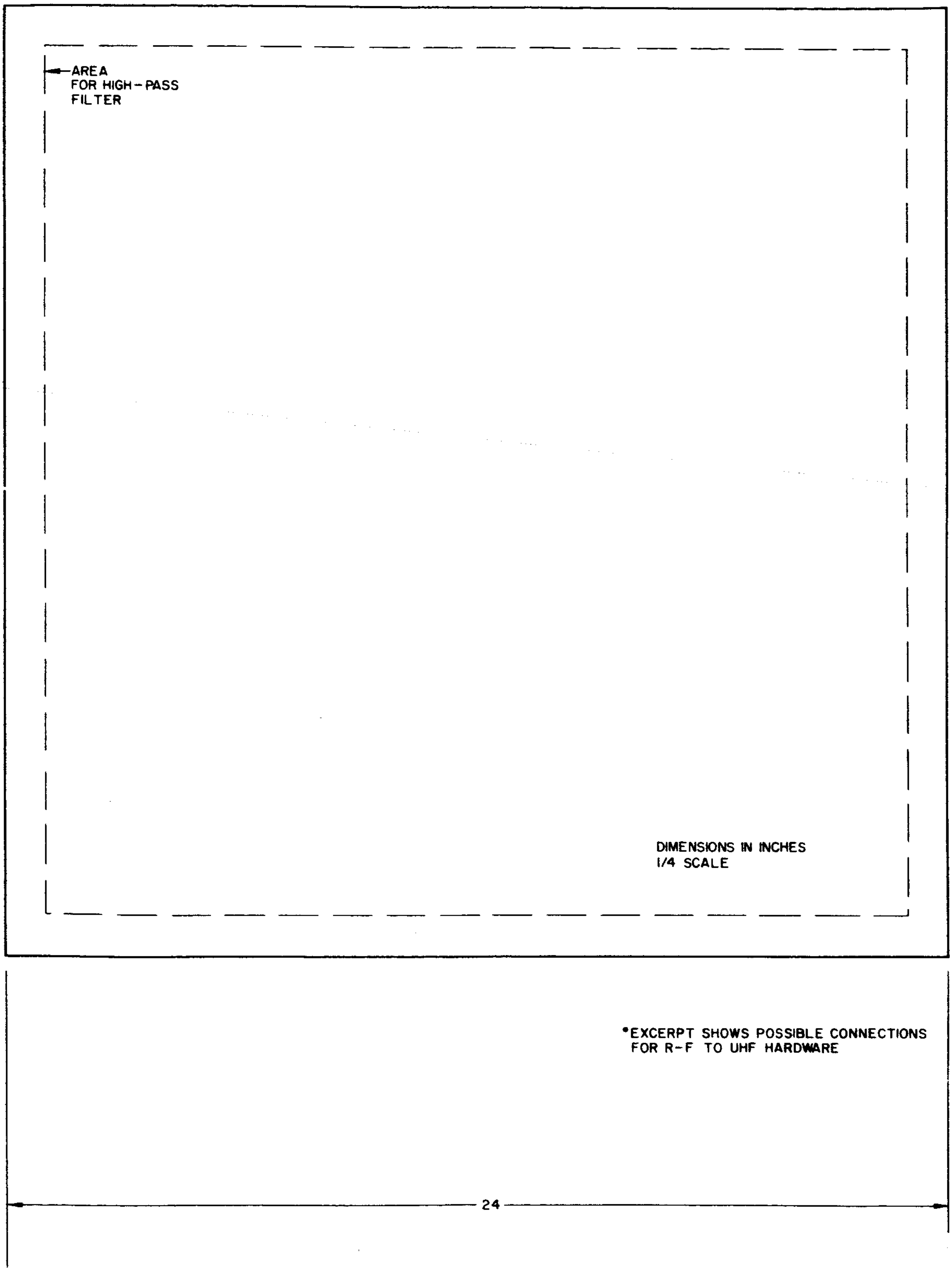
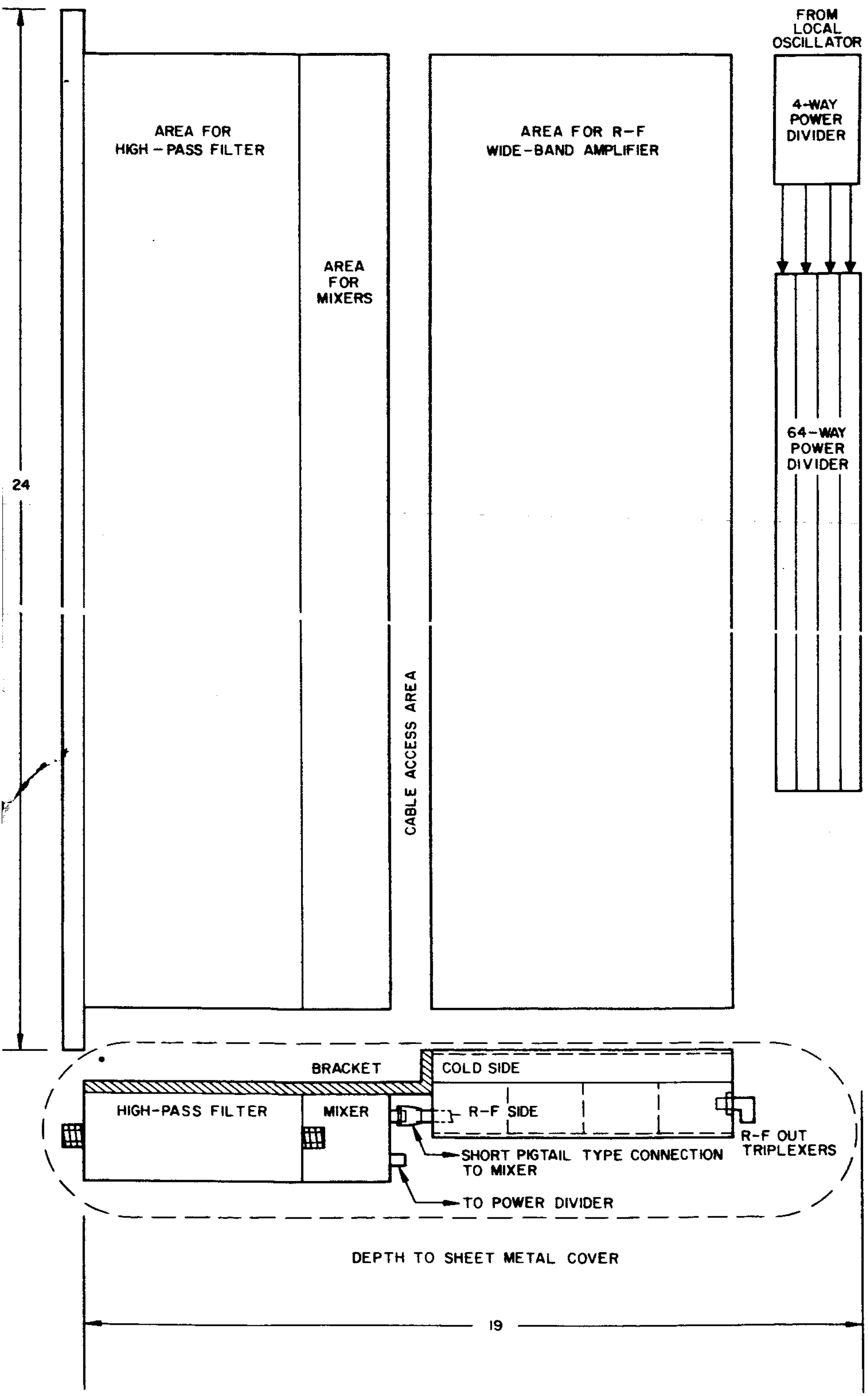


Figure 6-2. Receiving antenna layout for integrated system engineering model.



FROM LOCAL OSCILLATOR
4-WAY POWER DIVIDER
AREA FOR HIGH - PASS FILTER
AREA FOR MIXERS
AREA FOR R-F WIDE-BAND AMPLIFIER
64-WAY POWER DIVIDER
24
CABLE ACCESS AREA
BRACKET
COLD SIDE
HIGH-PASS FILTER
MIXER
R-F SIDE
R-F OUT
TRIPLEXERS
SHORT PIGTAIL TYPE CONNECTION TO MIXER
TO POWER DIVIDER
DEPTH TO SHEET METAL COVER
19

2

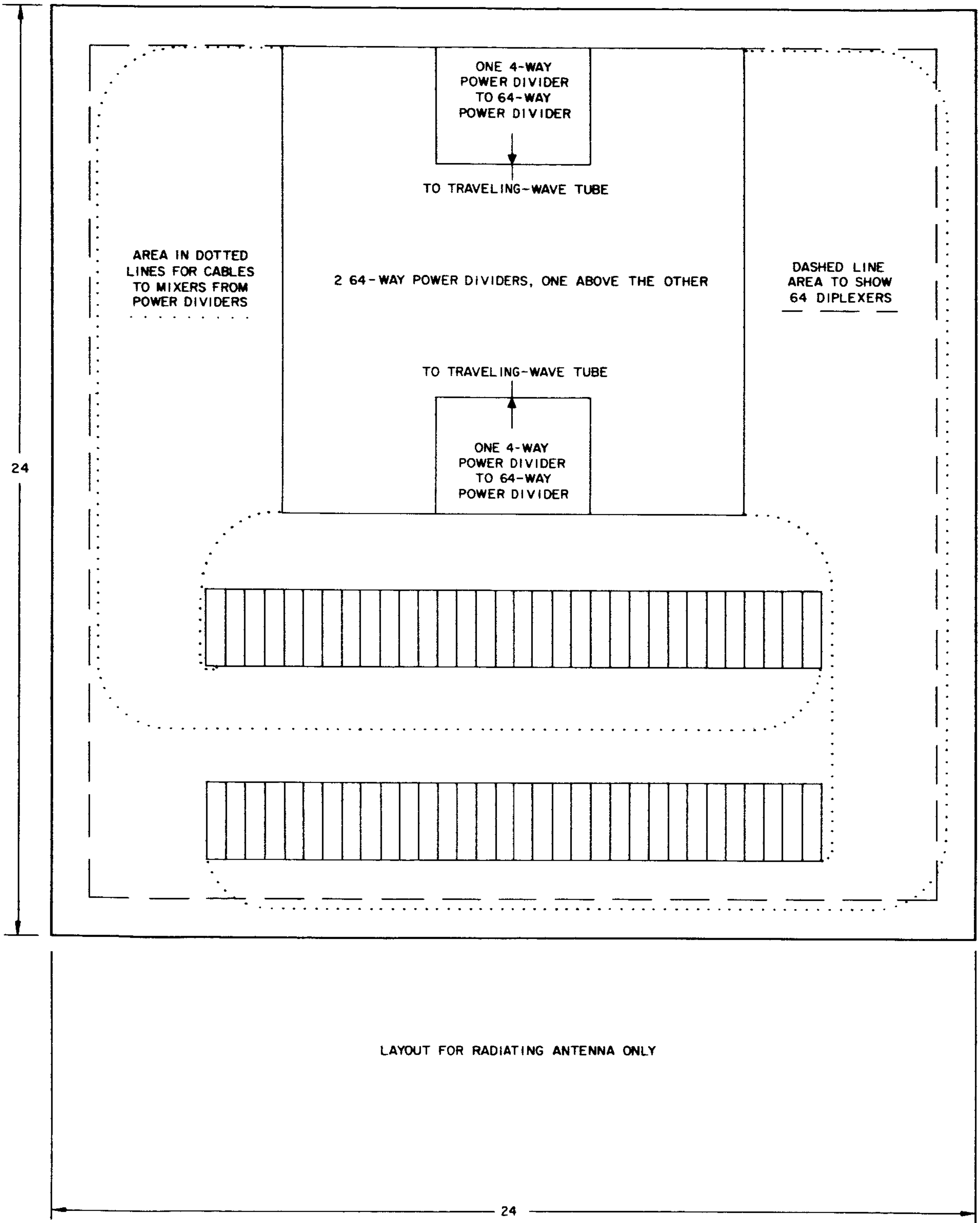
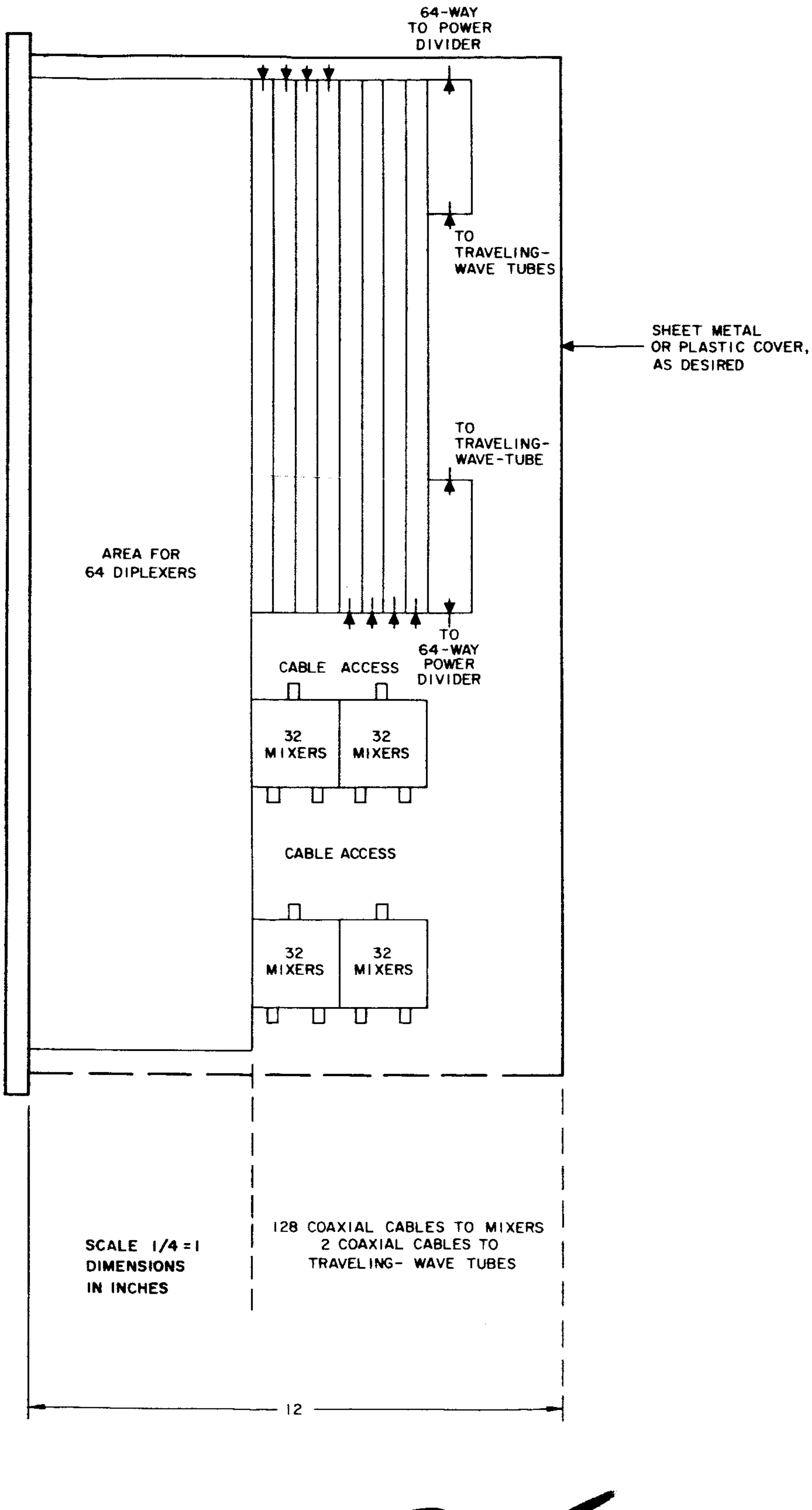


Figure 6-3. Transmitting antenna layout for integrated system engineering model.


64-WAY
TO POWER
DIVIDER
TO
TRAVELING-
WAVE TUBES
SHEET METAL
OR PLASTIC COVER,
AS DESIRED
TO
TRAVELING-
WAVE-TUBE
AREA FOR
64 DIPLEXERS
TO
64-WAY
POWER
DIVIDER
CABLE ACCESS
32
MIXERS
32
MIXERS
CABLE ACCESS
32
MIXERS
32
MIXERS
SCALE 1/4 = 1
DIMENSIONS
IN INCHES
128 COAXIAL CABLES TO MIXERS
2 COAXIAL CABLES TO
TRAVELING- WAVE TUBES
12

2

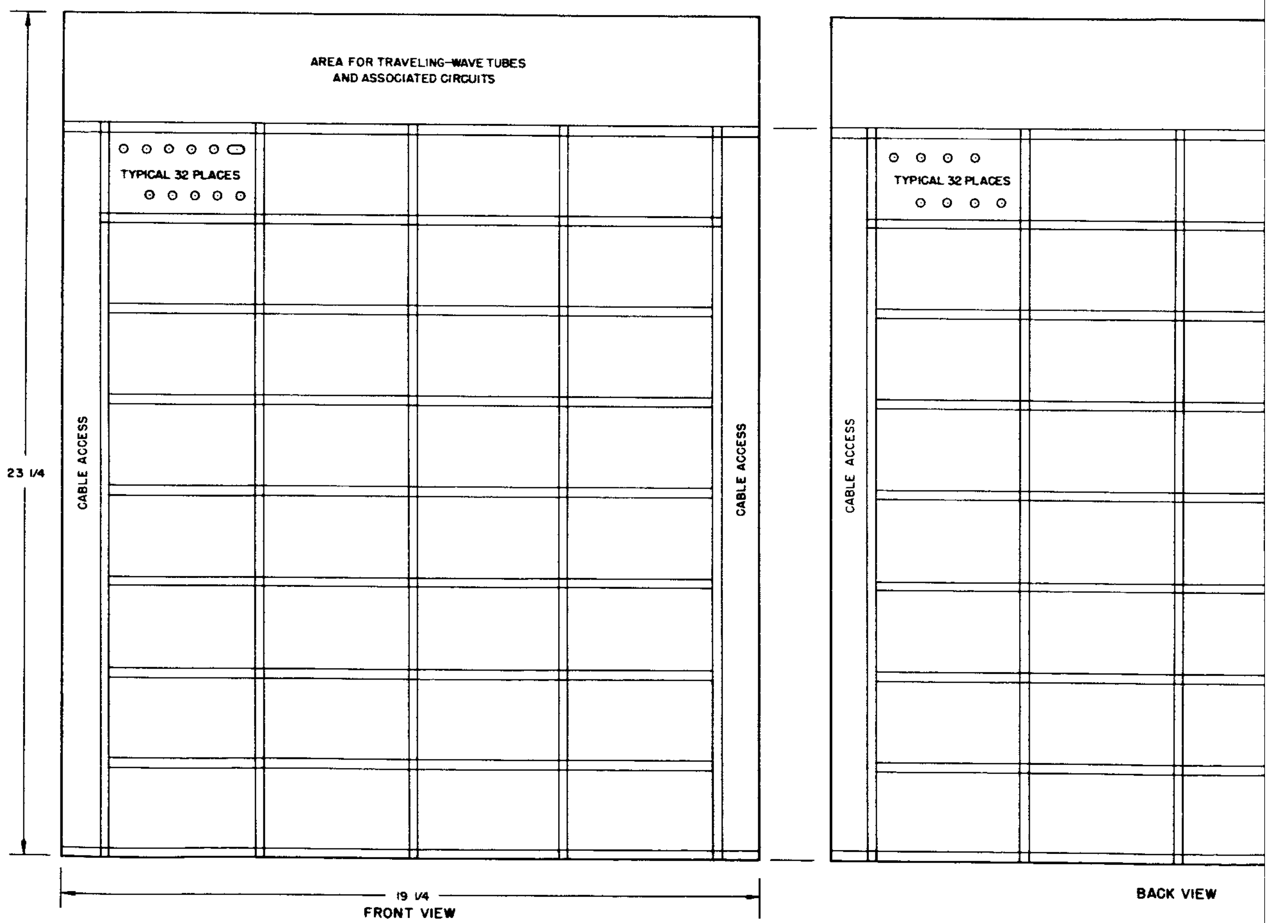
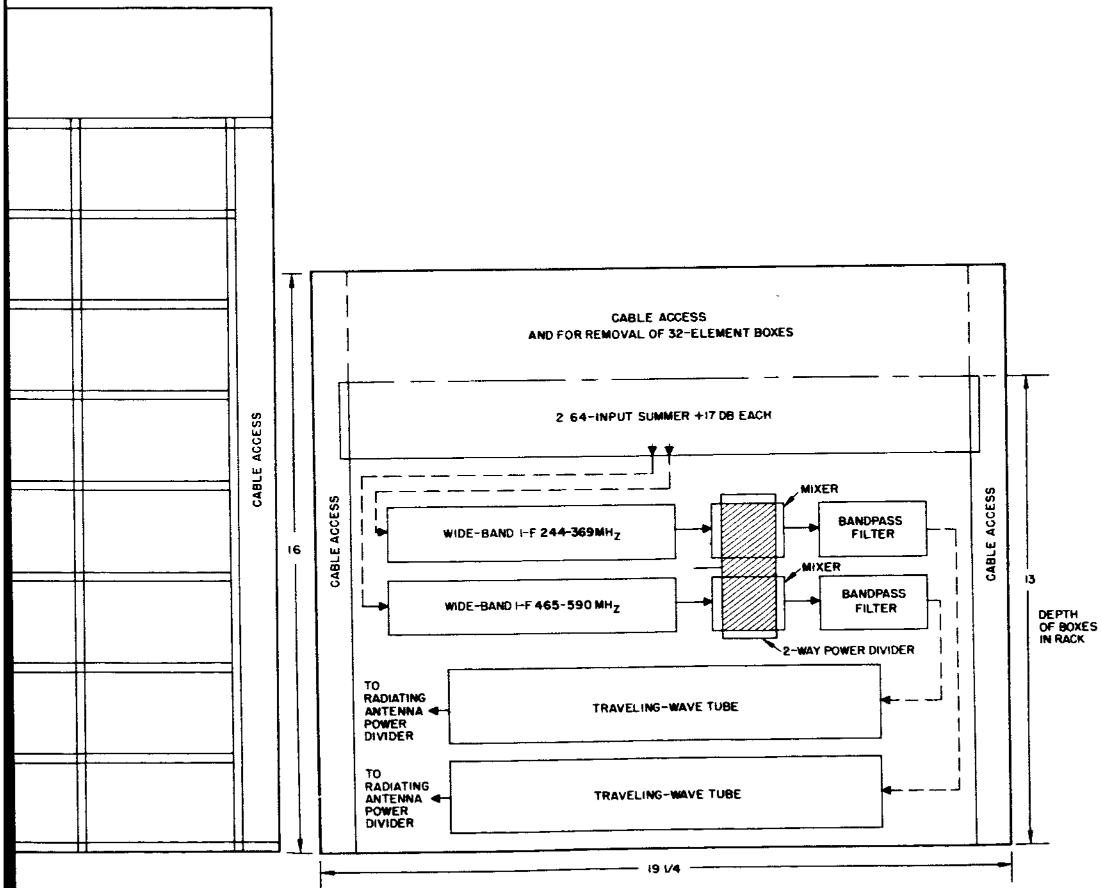


Figure 6-4. Main 32-element circuit rack layout showing traveling-wave tubes and associated circuitry for integrated system engineering model.


CABLE ACCESS
16
CABLE ACCESS
AND FOR REMOVAL OF 32-ELEMENT BOXES
2 64-INPUT SUMMER +17 DB EACH
CABLE ACCESS
WIDE-BAND I-F 244-369MHz
MIXER
BANDPASS FILTER
WIDE-BAND I-F 465-590 MHz
MIXER
BANDPASS FILTER
2-WAY POWER DIVIDER
CABLE ACCESS
13
DEPTH OF BOXES IN RACK
TO RADIATING ANTENNA POWER DIVIDER
TRAVELING-WAVE TUBE
TO RADIATING ANTENNA POWER DIVIDER
TRAVELING-WAVE TUBE
19 1/4

TOP VIEW

2

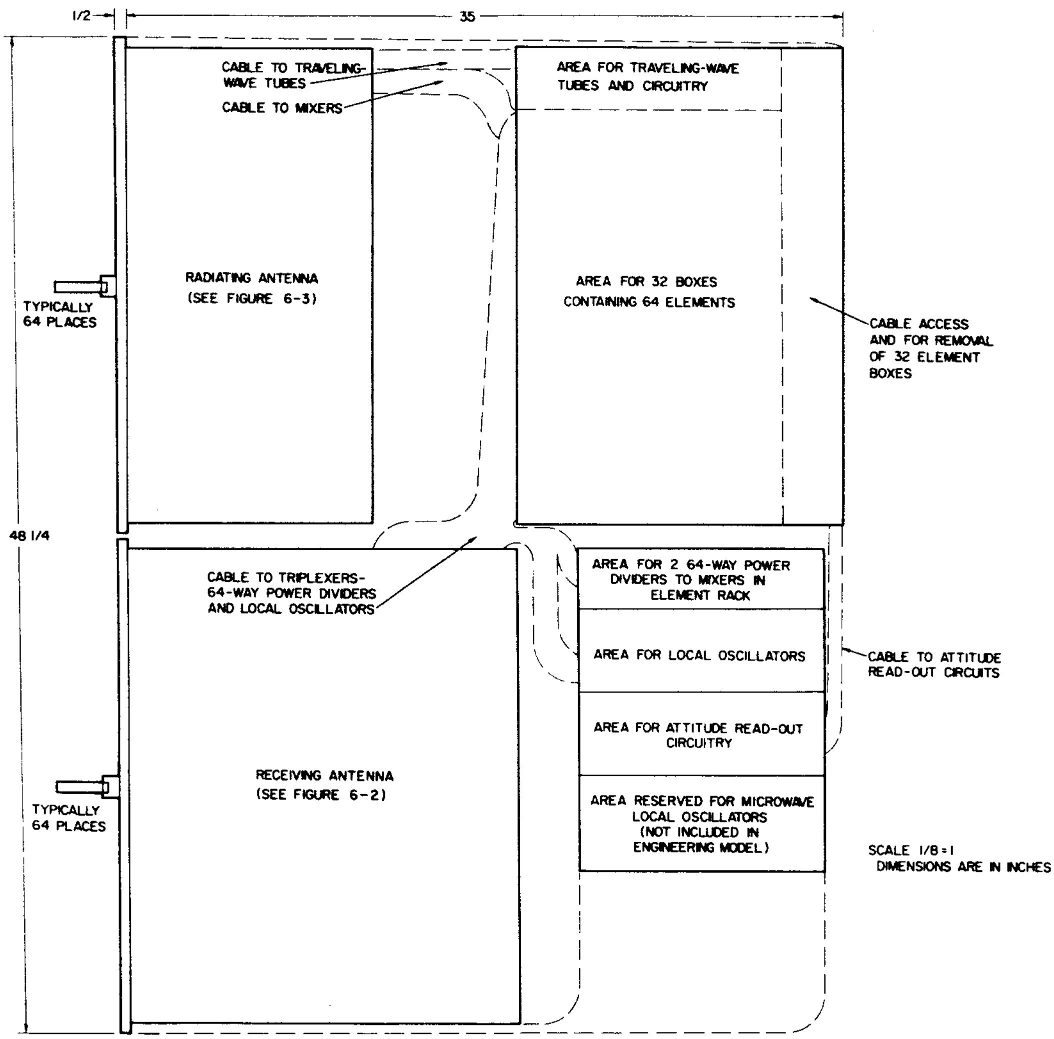


Figure 6-5. Integrated system side-view layout showing basic units with circuitry, cable connections, and cable access ways.

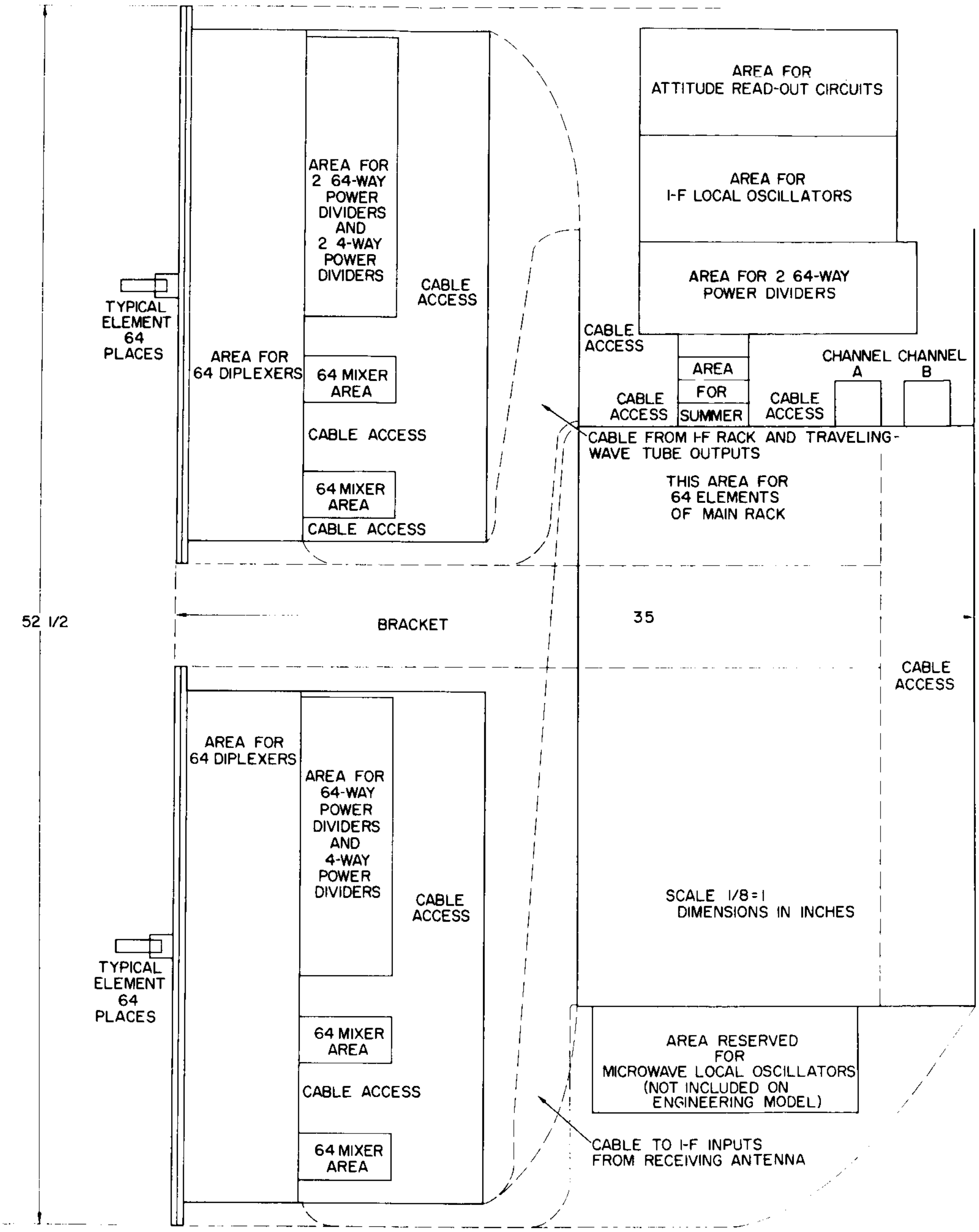


Figure 6-6. Dual-channel alternative system configuration layout.

the traveling-wave tubes and associated circuitry is indicated in Figure 6-4.

6.3.2 Receiving Antenna Layout

The high-pass filters are to be bolted to the antenna ground-plane at a location corresponding to an antenna element so that the element can be screwed directly into the input of the filter. The mixers will be bolted to the high-pass filters, and a pig-tail output terminal will be provided for connections to the preamplifiers. This arrangement will provide flexibility for any antenna element configuration on a flat surface.

A lightweight structure will be used at the output of the preamplifiers to provide strength and a plane for i-f and r-f connectors. The 64-way power divider will be located behind the preamplifiers at a location to minimize cable lengths. Sixty-four cables carrying i-f signals will connect the receiving antenna to the i-f electronics unit.

Layout of the receiving antenna can be seen in Figure 6-2.

6.3.3 Transmitting Antenna

The transmitter diplexers will be bolted to the antenna ground-plane and their input connectors will provide the antenna elements with mounting terminals. The mixers will be mounted in four rows of thirty-two each on a lightweight structure, along with the two 64-way power dividers required for the transmitting array. This approach requires more cabling but will conserve space needed for the 130 cables carrying i-f signals from the electronics unit.

Layout of the transmitting antenna can be seen in Figure 6-3.

6.3.4 Trade-offs

Inquiries are being made with respect to integrating the microwave circuitry to reduce weight, volume, and interconnections. It is possible that the mixers and 64-way power dividers can be integrated, with the high-pass filters and diplexers left as separate units. This arrangement may be necessary because the high insertion loss of strip transmission line filters may degrade system performance too much.

Unit			Total Weight (pounds)
Electronics			71.2
structure		80 oz	
i-f module	28 oz x 32	896 oz	
64-input summer	12 oz x 2	24 oz	
64-way divider	12 oz x 2	24 oz	
bandpass filter	1.5 oz x 2	3 oz	
mixer	2 oz x 2	4 oz	
2-way power divider		2 oz	
i-f local oscillator	5 oz x 2	10 oz	
attitude read-out		32 oz	
TWT	32 oz x 2	64 oz	
		1139 oz	
Transmitting Antenna			51.5
antenna structure and elements		192 oz	
mixers	2 oz x 128	256 oz	
diplexers	4 oz x 64	256 oz	
64-way power divider	12 oz x 2	24 oz	
coaxial cabling		96 oz	
		824 oz	

Table 6-1. Estimated unit weights of engineering model.

Unit			Total Weight (pounds)
Receiving Antenna			38.7
antenna structure and elements		192 oz	
high pass filters	2 oz x 64	128 oz	
mixers	2 oz x 64	128 oz	
64-way power divider		12 oz	
wideband i-f preamp	1.5 oz x 64	96 oz	
coaxial cabling		64 oz	
		620 oz	
Interconnection Cables			13.5
power cables	0.75 oz x 32	24 oz	
triaxial cables	0.42 oz x 458	192 oz.	
		216 oz	
System Weight Total			174.9

Table 6-1 (continued). Estimated unit weights of engineering model.

Added Components	Weight (pounds)
32 i-f modules 1.75 lb each	56
64 mixers 2 oz each	8
64 diplexers 4 oz each	16
1 64-way power divider	0.75
additional coaxial cables	4
	88.75 pounds

Subtracted Components	Weight (pounds)
64 high-pass filters 2 oz each	8
64 wide-band preamplifiers 1.5 oz each (included in 1-f modules)	6
	14 pounds
Net increase in weight	74.75 pounds
Total weight of alternate system	250 pounds

Table 6-2. Differential weight of alternative system.